Vol 7 No. 5, pp. 129-160 MAY 1942

Philips Technical Review

DEALING WITH TECHNICAL PROBLEMS
RELATING TO THE PRODUCTS, PROCESSES AND INVESTIGATIONS OF
N.V. PHILIPS' GLOEILAMPENFABRIEKEN
EDITED BY THE RESEARCH LABORATORY OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN, EINDHOVEN, HOLLAND

A NINE-KILOWATT EXPERIMENTAL TELEVISION TRANSMITTER

by M. VAN DE BEEK.

621.397.6

While in the case of a broadcasting transmitter the modulation spectrum has a width of only a few kc/sec, the spectrum of the video-frequency signals with which the carrier wave of a television transmitter is modulated occupies a region several Mc/sec in width. This fact, as well as the related fact that a television transmitter must work on very short wave lengths, causes the transmitting apparatus of a television transmitter to differ considerably in many respects from that of a broadcasting transmitter. The description here given of the powerful experimental television transmitter in Eindhoven attempts to present these differences clearly. Following a short description of the excitation of the carrier-wave frequency (wavelength 7 m) a detailed discussion is given of the method of modulation, the dimensions and construction of the output stage, which contains two water-cooled pentodes type PAW 12/15 and furnishes a maximum power of 9 kW to the transmitting aerial, as well as several particulars of the modulator.

The development of television technique has from the very beginning aroused considerable interest. It has been particularly the methods of recording and reproducing the pictures which have attracted the most attention and which have been dealt with in many technical and popular treatises. The problem of transmitting the electrical signals into which the picture is translated was usually less well described, since in principle the wireless transmission of signals was already familiar from broadcasting and therefore provided no new technical sensations. In this periodical also the subject has only been mentioned incidentally, while studio equipments and receivers have repeatedly been dealt with elaborately [1]. Notwithstanding all this, the actual transmitting apparatus of a television installation has its own specific problems which arise from the great width of the frequency band which must be faithfully transmitted as far as amplitude and phase are concerned, and from the short wave length connected with the former on which the transmitter must work. In the following article, in which we shall describe the experimental television transmitter in Eindhoven, this will become quite clear.

The transmitter in question originated from the experimental television transmitter which was developed here in 1935 [2] and which furnished an output of several hundred watts. The output has been increased to approximately 9 kW (maximum output radiated by the aerial).

Survey of the complete transmitter

In *fig. 1* the electrical structure of the whole is represented in the form of a block diagram. The carrier-wave frequency is derived from the oscillation of a plate of quartz crystal which vibrates at a frequency of 5.4 Mc/sec (wave.

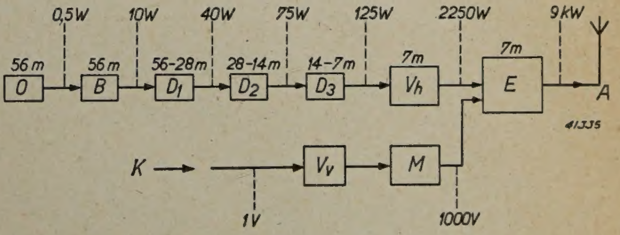

Fig. 1. Block diagram of the television transmitter. *O* crystal oscillator, *B* buffer stage, $D_1D_2D_3$ frequency doubling stages, V_h high-frequency amplifier stage, *E* output stage, *K* cable from studio, V_v video-frequency amplifier, *M* modulator, *A* aerial.

---

[1] Philips Techn. Rev. 1, 16, 1936 (An experimental television transmitter and receiver); 1, 321, 1936 (Television); 2, 33, 1937 (A television receiver); 2, 72, 1937 (Television system with Nipkow disc); 2, 249, 1937 (The enlarged projection of television pictures); 3, 1, 1938 (A transportable television installation); 3, 285, 1938 (Television with Nipkow disc and interlaced scanning); 4, 42, 1939 (The Nipkow disc); 4, 342, 1939 (Television receivers).

[2] J. van der Mark, An experimental television transmitter and receiver, Philips Techn. Rev. 1 16, 1936.

length 56 m). In a number of amplifier stages, which at the same time serve partially as frequency doublers, the high-frequency output is increased to about 100 W, at the desired wavelength of 7 m. The following amplifier stage furnishes the power necessary for the excitation of the large water-cooled transmitter valves which are included in the output stage. This output stage, in which at the same time the video-frequency signals furnished by the studio are modulated on the carrier wave, delivers to the transmitting aerial a maximum energy of 9 kW.

Oscillator and frequency-doubling stages

By the employment of a quartz plate cut from a quartz crystal along a so-called temperature-insensitive plane, a very constant carrier-wave frequency is obtained without it being necessary to place the oscillator stage in a thermostat. There are two reasons why the excitation of the desired carrier-wave frequency is begun with a quartz plate which vibrates with a much lower frequency. In the first place it is practically impossible to make quartz plates for wavelengths shorter than about 20 m, since they become too thin. In the second place with an initial wavelength which is not too short it is much easier to attain a satisfactory amplification, so that fewer amplifier stages are needed than when the oscillator stage already operates on 7 m.

In order to prevent loading variations in the last stages, due for instance to the modulation, from reacting on the oscillator stage (and thereby on the carrier-wave frequency), the first amplifier stage is constructed as a buffer stage, *i.e.* the control grid of the amplifier valve in question is given such a high negative bias as to prevent flow of grid current, so that there is practically no load on the oscillator stage. This is in contrast to the ordinary situation in transmitter stages, where in order to attain a reasonable efficiency the grid voltage is chosen so that grid current shall flow.

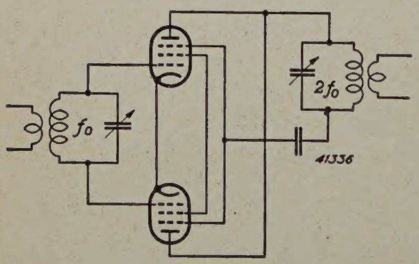

Fig. 2. Diagram of the connections of the doubling stages. The two valves are excited in opposite phase; the anodes are connected in parallel. The anode circuit is tuned to a frequency twice as high as that of the grid circuit. In order to give the free end of the anode circuit as constant a potential as possible, *i.e.* one which varies as little as possible at a high frequency, with the very short waves used here a connection with earth is not suitable, since the long connection wires may have considerable impedances for these waves. Therefore the anode circuit is connected here with the mutual connection of the screen grids of the two valves, which, with the valves used, leads also to a very simple structural solution, since anode and screen grid leads are placed at the top of the valve envelope.

Each frequency-doubling stage contains two valves which are connected in the manner shown in *fig. 2.* Each valve, since it operates in Class C arrangement, only furnishes anode current during less than half a period of the excitation voltage. Since the control grids are excited in opposite phase, where as the anodes are connected in parallel, an anode current of the form shown in *fig. 3* is obtained. The fundamental frequency of this current,, which is filtered out by the tuned anode circuit, is double the frequency of the excitation voltage. In this way the desired doubling of the frequency is obtained with the relatively high efficiency of about 55 percent. A pleasing property of the chosen system of amplification with frequency doubling is that, due to the difference between the frequencies of input and output voltage, no reaction is possible and therefore no self generation of the frequency-doubling stages can occur.

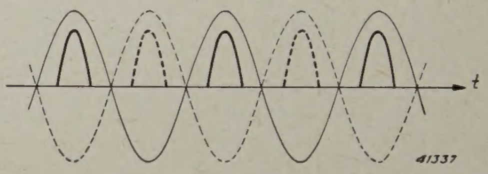

Fig. 3. Grid voltage (thin line) and anode current (heavy line) of the frequency-doubling stage. The full drawn lines are for one valve, the broken lines for the other. The total anode current has a fundamental frequency twice as high as the grid voltage.

Preliminary considerations in the design of the output stage

In order to discuss the connections and dimensions of the other transmitter stages, particularly the output stage, it is desirable first to go back to the video-frequency signal as it is sent from the studio to the transmitter.

When n pictures per second are transmitted, each having N lines, while the ratio of width to height of picture is b/h, the lowest modulation frequency 2) occurring is $f_{\min} = n$ c/sec, the highest $f_{\max} = {}^1/_2\, n . N^2\, b/h$ c/sec. In our case $n = 25$, $N = 405$ and $b/h = 1.2$, so that one must count on modulation frequencies between 25 c/sec and 2.5 Mc/sec. When this video-frequency spectrum is modulated on a carrier wave there occur, in addition to the carrier-wave frequency, two side bands, each 2.5 Mc/sec in width (*fig. 4*). From the point in the transmitter where the carrier wave is modulated up to and including the aerial the connections must therefore have a transmitting region (band width) of at least 5 Mc/sec. This condition is made more imperative by the fact that all the vibrations within the frequency band mentioned must not only be transmitted in the correct amplitude relations, but also, in contrast to broadcasting transmitters, in a certain relative phase: the phase shift must vary linearly with the frequency so, that no relative displacement of points in the picture and thus distortion may occur. Since in every case such a linear relation between phase and frequency can only be obtained for phase angles which are not too large, the last requirement mentioned comes down to the condition that the phase rotations must only be very small [3]).

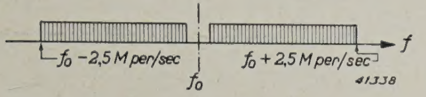

Fig. 4. Frequency spectrum of the carrier wave modulated with the video-frequency signal.

What consequences has this for the connections?

In the first place the requirement relating to the phase rotation implies that no band-pass filters can be used for the realization of the wide transmission region, since with such filters large phase shifts always occur. Recourse must therefore be had to single circuits, which may be represented by the diagram of *fig. 5*: a loss-free self-induction L and capacity C, with a loading resistance R. This loaded circuit is connected with a source of constant alternating current i (a pentode for instance). The impedance $Z = V/i$ of the circuit, which in this case represents the behaviour of the voltage V on the circuit, is drawn in *fig. 6* (for certaln values of L, C and R) according to magnitude $|Z|$ and phase φ, as a function of the frequency. If for the outermost side-band frequencies we allow a fall in the amplitude from 1 to $1/\sqrt{2}$, a frequency region of the width

$$2\,\Delta f = \frac{1}{2\pi\, RC} \quad \ldots\ldots\ldots (1)$$

around the resonance frequency $f_0 = {}^1/_{2\pi}\,\sqrt{LC}$ is found to be useable. In this frequency region, according to fig. 6*b*, the phase remains below 45° and varies practically linearly with the frequency.

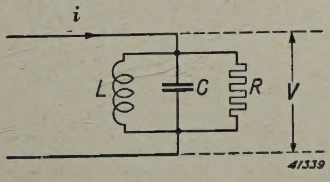

Fig. 5. Diagram of loaded tuned circuit.

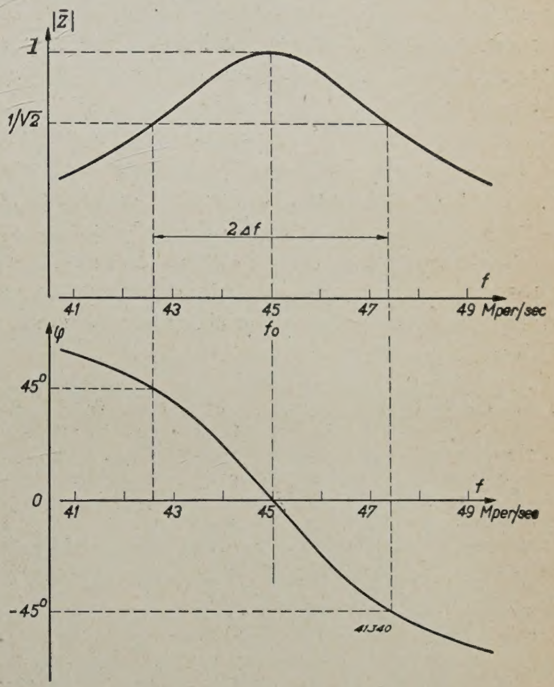

Fig. 6. *a*) Absolute value $|Z|$ and *b*) phase φ of the impedance of a tuned circuit according to fig. 5, as a function of the frequency f. For the circuit elements the following values are here assumed: $L = 0.395$ μH, $C = 31.8$ pF, $R = 1000$ ohms.

In order to give the circuit the desired great band width of $2\Delta f = 5$ Mc/sec, according to eq. (1), R and/or C must be made very small (a decrease in C must be accompanied by an increase in L, in order that the resonance frequency may remain in its place). Now C cannot be made arbitrarily small, because even when no capacity at all is introduced into the circuits the capacities of the transmitting valves in parallel always act over them. In push-pull connections of two medium-sized valves there is in this way always a capacity of at least

---

[3]) A similar requirement is encountered in carrier-wave telephony, where wide frequency bands are also used; see for example F. de Fremery and G. J. Levenbach, Carrier-wave telephony on coil-loaded cables, Philips Techn. Rev. 4, 20, 1939. If the requirement of linearity is satisfied no "phase distortion" occurs, but the signal has a certain transit time which may sometimes be quite long. While in carrier-wave telephony certain limits must also be set to this, the finite transit time is no objection at all in television.

30 pF. The result is that in order to obtain the required band width the resistance R may not be made larger than $^1/_4\pi \Delta f C \approx 1000$ ohms. This is very unfortunate, since the power output of the transmitter valve, of which the circuit forms the load on the anode, is proportional to R within certain limits.

If there are several modulated stages the total frequency characteristic of the transmitter is obtained by multiplying the amplification factors of all the modulated stages with each other at each frequency, and adding up the phase rotations. Therefore, for each separate stage, a stricter criterion than the above must be chosen for the useable band width, so that for each of these stages the anode impedance must again be considerably lower. It is therefore clear that a television transmitter should always be modulated in the output stage. For a given output of the transmitter one then obtains on the one hand a higher efficiency and output from each of the preliminary stages, so that the number of such stages may be smaller, while on the other hand the output stage may still retain a relatively high anode impedance and thus this stage, which is the most expensive one as far as valve and current consumption are concerned, can operate with the greatest possible efficiency.

Since the quality factor Q of a resonance circuit may be written as $Q = f_0/2\Delta f$, a great band width is always accompanied by a poor quality factor (poor building up), so that doubt may arise whether the introduction of the self-induction for the tuning of the anode circuits is actually worth the trouble. The tuned circuits do, however, justify their use for the suppression of the harmonics of the carrier-wave frequency — in our case, where $f_0 = 43.2$ Mc/sec and $2\Delta f = 5$ Mc/sec, the second harmonic is attenuated to about 1/20 and the third to about 1/30 — and to prevent the video-frequency signals from reaching the aerial directly, which would cause radio interferences in a wide region around the television transmitter on all wavelengths above 120 m.

Let us now consider the choice of a modulator system. Should the carrier wave be modulated by variation of the anode voltage or of the grid voltage of the output stage? Grid voltage modulation is generally employed in television transmitters for the following reasons. In the first place, from the point of view of simplicity of construction, grid modulation is preferable, since with anode modulation a very large modulation power is needed because of the high anode voltages of the output stage. This leads to a large and expensive modulator for the necessary wide modulation spectrum. In the second place, and this is even more important, with grid modulation the synchronization impulses can be better transmitted. While all the gradations of brightness of the picture to be transmitted are transmitted by varying the carrier-wave amplitude between 30 and 100 percent, each synchronization impulse consists of a short-lived suppression of the carrier wave (*fig. 7*). With anode modulation the anode voltage may indeed become equal to zero during the synchronization impulses, but in spite of this, due to the induction effect, a certain anode alternating current continues to flow [4]) and there is still, therefore, some radiation by the aerial. With grid modulation, on the other hand, it is possible to make the control-grid voltage so strongly negative at those impulses that the output valve passes absolutely no anode alternating current. Sharper synchronization signals are thus obtained and thereby a more reliable functioning of the receiving sets.

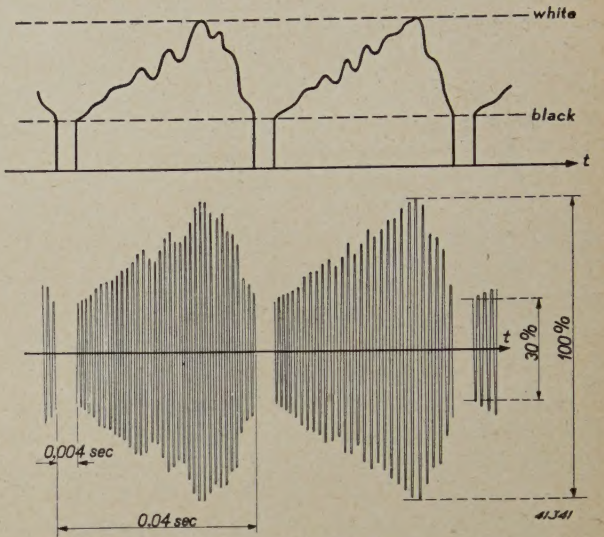

Fig. 7. *a*) Variation with time of the video-frequency signal, schematically. *b*) Variation with time of the high-frequency signal emitted by the television transmitter. (The line synchronization signals are omitted for the sake of simplicity).

It is in fact less easy with control grid modulation than with other methods of modulation to realize a strictly linear modulation [5]) (*i.e.* a linear relation between carrier-wave amplitude and momentary value of the modulation voltage). In the case of a television transmitter, however, this is much less important than in the case of a broadcasting transmitter, since the eye is much less sensitive to non-linear distortion than the ear. It is a familiar fact that large distortions in the rendering of brightness gradations, for instance in photography, can be tolerated without producing an "unnatural" impression on the eye.

---

[4]) See Philips Techn. Rev. **1**, 176, 1936.

[5]) See J. P. Heyboer, Five-electrode transmitting valves, Philips Techn. Rev. **2**, 257, 1937.

Dimensions of the output stage

The output stage contains two water-cooled pentodes, type PAW 12/15, in push-pull connection (these are the largest pentodes made). As was explained above, the capacity of the anode circuit must be made as small as possible. In anticipation of the description of the construction to be given below, it may be stated that this capacity could be limited to 32 pF. According to eq. (1) an anode loading with an impedance of about 1000 ohms is then permissible to obtain the desired band width.

In order to develop as much high-frequency energy as possible in this anode impedance, which is very low compared with broadcasting transmitters, the anode current must be modulated as much as possible, preferably to saturation. What anode DC voltage, excitation voltage and modulation voltage must be chosen for this, and what output can be obtained?

The valve PAW 12/15, with a normal tungsten filament for 22 V, 80 A, has an emission of 11 A. This, however, is not fully available, since, in addition to the anode, the control grid and the screen grid also require their part of it. If on the basis of experience we reserve 1.75 A for the control grid current and 0.75 A for the control grid peak current, there remains available for the anode peak current I 8.5 A.

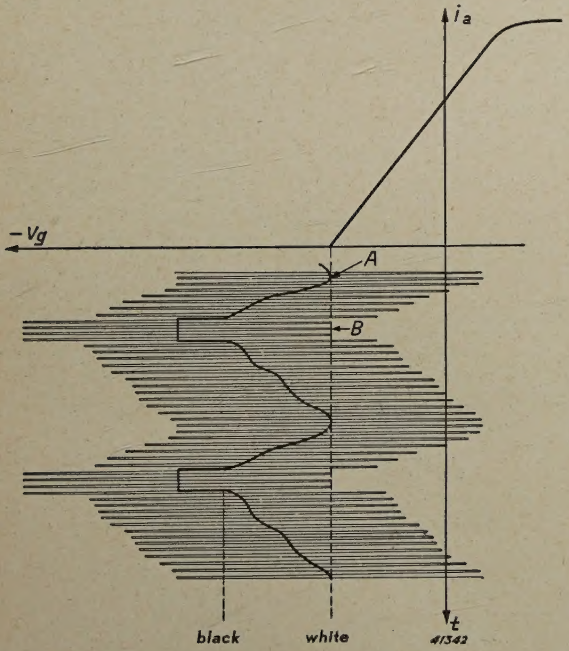

Fig. 8. Variation of the grid voltage of the output stage with control grid modulation. High-frequency excitation voltage and video-frequency modulation voltage are superposed in such a way that at the peak adjustment (completely white picture, point A) the valve just operates in class B (grid bias equal to overbiasing voltage), and that the carrier wave is suppressed (peak of excitation still below overbiasing voltage) during the synchronization impulses (point B).

In order to obtain as linear a modulation as possible in both valves in push-pull connection, the control grid modulation is so effected that upon adjustment to the largest carrier-wave amplitude (completely white picture) the valves just operate in "class B", *i.e.* the control grid DC voltage is equal to the overbiasing voltage, see *fig. 8.* In this way at this peak adjustment (A) the anode current takes on the form indicated in *fig. 9.* The first harmonic, with which we are concerned, has, as the Fourier series shows, the amplitude $I/2 = 4.25$ A. Now the action of the push-pull connections may be conceived as if each of the two valves were loaded with half of the anode impedance, *i.e.* 500 ohms. The power contributed to this by the anode alternating current amounts to $\frac{1}{2} \cdot 4.25^2 \cdot 500 = 4.5$ kW. Thus at the peak adjustment the two valves give a total output of 9 kW. Since only in the case of a completely white picture do the valves operate at this peak adjustment, and with entirely black picture the carrier-wave amplitude is adjusted to 30 percent of the maximum, the average output radiated amounts to about 3 kW.

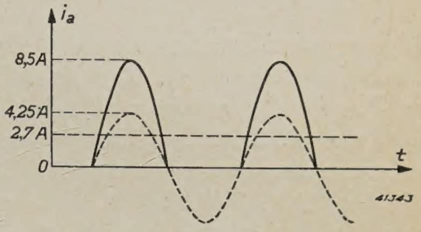

Fig. 9. Form of the anode current in one valve of the output stage, at the peak adjustment. The available emission permits a peak value of 8.5 A; first harmonic 4.25 A, average value (anode direct current) 2.7 A.

It is further evident from the above that the anode AC voltage at the peak adjustment amounts to $4.25 \times 500 = 2125$ V. Since it is known from the static I_a-V_a characteristics of the valve PAW 12/15 that the momentary value of the anode voltage may not be lower than about 2600 V, the anode DC voltage must be taken approximately equal to 4750 V. Since for a current of the form shown in fig. 9 the anode DC $I/\pi = 2.7$ A, a DC power of $4750 \times 2.7 = 12.8$ kW is applied to each valve. The difference between this power applied and the output, *i.e.* 12.8—4.5 = 8.3 kW, is dissipated in the valve. The permissible dissipation (12 kW) is thereby by no means exceeded. The efficiency of the anode at the peak adjustment is $100 \times 4.5/12.8 = 39$ percent.

The fact that the output of the valve in our case is limited by the emission of the filament, and not, as is often the case in broadcasting transmitters, by the permissible dissipation, is directly connected with the great band width required, and the small anode impedance thus necessary. With larger anode impedance the calculated current would result in a higher anode AC voltage, as a result of which the anode DC voltage would also have to be chosen higher. The power applied would then increase, and it would increase more rapidly than the power output, so that also the dissipation would quickly approach the permissible limit and a limitation of the anode current would finally become necessary. At the same time the voltage modulation, which in our case is only 2125/4750 ≈ 0.45, then approaches the more normal values in the neighbourhood of unity (see the article referred to in footnote [5]).

Let us now calculate the required excitation power of the output stage. The average slope (static and dynamic) of the valve PAW 12/15 is about 6 mA/V. Thus for the two valves in push-pull connection we need a grid AC voltage with the peak value 8.5 A : 6 mA/V = 1400 V. In modulation the total grid voltage will vary as in fig. 8, from which with the help of the characteristics it may be deduced that the average grid current at the peak adjustment will amount to about 300 mA. The theoretically necessary excitation power for the two valves together is therefore 2 × 0.3 × 1400 = 850 watts. However, due to the finite transit times of the electrons [6]) an extra grid damping occurs, as a result of which the excitation power is increased to about 1125 W. Moreover, it is necessary that during the modulation when the grid current is varying, the excitation power should remain constant. Since this is only possible with not too great loading fluctuations of the stage which furnishes the excitation voltage, the latter must be given a constant initial load independent of the grid current of at least the same magnitude as the excitation power, by means of grid damping resistances. This stage must therefore be able to deliver 2250 W of high-frequency power. This is obtained by means of four valves type PB 3/800 two by two in push-pull connection, each of which can deliver 625 W on 6.5 m. This stage itself in turn requires 4 × 25 = 100 W modulating power, which is furnished by the preceding frequency doubling stage (see the diagram of fig. 2).

Construction of the output stage.

The capacity of the anode circuit of the output stage, when no extra capacity at all is added, consists not only of the unavoidable grid-anode capacity of the two valves PAW 12/15 connected in push-pull, but also of various parasitic capacities: that between the two water-cooled anode jackets of the valves, between the cooling jackets and earth, etc. (see *fig. 10*). By placing the valves as shown in *fig. 11* at a good distance from each other and

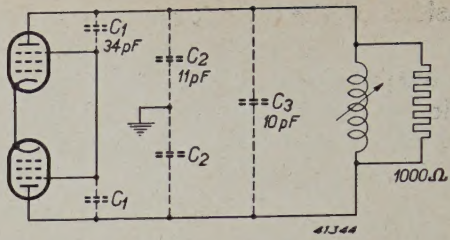

Fig. 10. The anode circuit of the output stage contains a capacity which is composed mainly of the anode-grid capacities C_1 of the two valves in series, the capacities C_2 between each anode cooling jacket and earth, in series, and the capacity C_3 between the two anode cooling jackets.

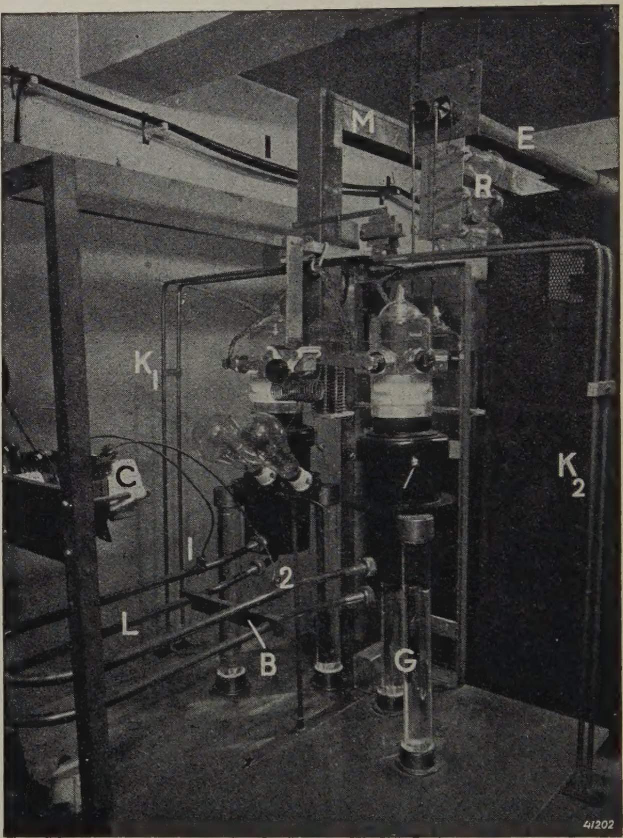

Fig. 11. Part of the output stage. The two water-cooled pentodes PAW 12/15 are mounted on glass columns G. L = Lecher system, which functions as self-induction and is tuned with the short-circuiting bridge B. At *1* and *2* the supply line is tapped off which conducts the high-frequency energy *via* the high-voltage coupling condensers C to the aerial. In order to be able to use the wires of the Lecher system, which are constructed in the form of tubes, for the inlet and outlet of the cooling water for the anode jackets, this system is built up of two identical parts in parallel. For the sake of simplicity the taps are placed on one part and the short-circuiting bridge on the other. K_1 and K_2 cathode supply, E lead for the excitation voltage with carbon filament lamps R serving as grid-damping resistance, M lead for the modulation voltage. Between the two valves may be seen the elements for the series tuning of control grid and screen grid circuit.

---

[6]) See for example the article referred to in footnote[4]); also Philips Techn. Rev. **6**, 129, 1941, footnote[10]).

at a generous height above the earthed base plate by means of long glass columns, these capacities could be reduced to the values given in fig. 10 and in this way the total circuit capacity to the value of 32 pF already mentioned.

The self-induction of about 0.4 μH required for the tuning to the desired carrier-wave frequency (43.2 Mc/sec) is realized by a short Lecher system connected with the anodes and tuned by means of a short-circuiting bridge [7]). Because of the fact that the distance apart of the Lecher wires must be sufficiently small, the two valves could not be placed any farther apart. The Lecher wires are in the form of tubes which serve at the same time for the inlet and outlet tubes for the cooling water for the anode jackets.

The load on the output stage is formed by the impedance of the transmitting aerial, which is coupled with the anode circuit *via* a supply line. Since the supply line is terminated at the aerial end by its wave resistance in order to avoid reflections, considered from the anode circuit end, it also has an impedance equal to the wave resistance, which amounts to 425 ohms. Since, however, according to the above, the circuit may be loaded with 1000 ohms, the supply line is connected by a "tap" on the Lecher system. In order to prevent the high DC voltage of the anode from reaching the aerial, two mica coupling condensers are inserted in the supply line. These details are represented diagrammatically in *fig. 12*; the practical construction is shown in fig. 11.

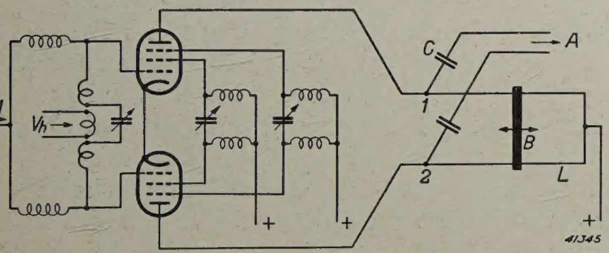

Fig. 12. Diagram of the practical construction of the output stage. At M the modulation voltage is applied, at V_h the excitation voltage. In all three grid circuits series tuning is introduced. L Lecher system, B short-circuiting bridge, C coupling condensers in the supply line to the aerial A.

Further particulars of the connections are the following. The two screen grids of the valves in push-pull connection are not connected directly with each other, but between them a tuning arrangement in series is introduced (see fig. 12). This was necessary since otherwise, due to the impedance of the external and internal connection wires, which cannot be ignored at these short wavelengths, high-frequency voltage still continued to act on the screen grids, so that they would not have fulfilled their purpose. With the control grid circuit the case was nearly the same. The internal capacities of this circuit are so large that even a direct connection of the grid terminals still possesses too much self-induction to give the desired high resonance frequency. Here also tuning in series was applied, and in this case the circuit capacity was so reduced by a series condenser that the self-induction necessary for the tuning took on a reasonably large value, so that a coil could be used, which at the same time makes possible a satisfactory coupling with the preceding stage.

The filaments are fed with AC voltage from the mains. In order to limit the hum percentage in the output signal, however, for the feeding of the filaments of the two valves in push-pull connection two voltages with a relative phase shift of 90° were used. In fig. 11 the two pairs of supply lines for this cathode feed may be seen to the left and right of the valves.

Dimensions and connections of the modulator

As we have already seen above, the largest carrier-wave amplitude is obtained by making the average control grid voltage equal to the overbiasing voltage of the valves (point A in fig. 8), while for the synchronization impulses the carrier wave must be entirely suppressed by making the average control grid voltage sufficiently negative (point B in fig. 8). From this it follows that the peak value of the modulation voltage must be equal to the amplitude of the excitation voltage, for which we found a value of 1400 V.

In order to give the modulator the necessary band width (from 25 to $f_{\max} = 2.5 \cdot 10^6$ c/sec) attention must again be devoted to a relation similar to eq. (1), *i. e.* the loading impedance R must be smaller than $^1/_{2\pi f_{\max} C}$[8]). The modulator is loaded by the grid circuit of the output stage, which circuit contains in parallel the grid-cathode capacities of the two valves in push-pull connection. In our case this is a total capacity or 116 pF. The anode impedance of the modulator may not therefore amount to more than $R = 10^{12}/116 \cdot 2\pi \cdot 2.5 \cdot 10^6 = 550$ ohms. Since with the prescribed voltage the power is inversely proportional to R, this makes it necessary to have a large video-frequency powe (more than 1500 W). This is obtained with two valves type PB 3/800 connected in parallel.

---

[7]) The fact that a Lecher system whose length is less than $^1/_4$ wavelength behaves in the main as a self-induction is explained in C. G. A. von Lindern and G. de Vries, Lecher systems, Philips Techn. Rev. 6, 240, 1941.

[8]) See for example the discussion of a broad-band amplifier in J. D. Veegens, A cathode-ray oscillograph, Philips Techn. Rev. 4, 198, 1939.

Since the video-frequency signal which comes from the studio amounts to only 1 volt across the wave resistance of 1000 ohms with which the connecting cable is terminated at the transmitter end, a number of amplifier stages must precede the modulator. In order to do with as few stages as possible, valves with the steepest possible slope were chosen, while in several stages two or even three valves were connected in parallel. The advantage in steepness gained by this means is of course partially compensated by the fact that the input and output capacities become correspondingly greater, so that the loading resistance of the stages in question had again to be reduced.

We shall not go into detail here about the further particulars of the pre-amplifier, such as measures for suppressing tendencies to generate, for improving the phase reliability at the low frequencies, for the suppression of interferences by mains voltage variations etc. One important difference, however, between the video-frequency modulator and the audio-frequency modulator in broadcasting transmitters will be discussed in somewhat more detail.

In contrast to the AC voltages which are used to transmit speech or music, the video-frequency AC voltages which are delivered by the iconoscope have an average value which differs essentially from zero. The magnitude of this average value (DC voltage component) is in fact a measure of the average brightness of the picture. Actually, therefore, the video-frequency amplifiers in the studio as well as in the transmitter should be constructed as DC amplifiers. There are, however, many objections to this in connection with the broad spectrum and the high power finally necessary. Therefore in the case of television transmitters the DC voltage component is usually filtered out in the studio amplifier, so that all the remaining amplifier stages can be constructed as AC amplifiers, and only at the input of the last amplifier stage (*i.e.* the modulator stage proper) is the DC voltage component applied again. The modulator stage itself is therefore constructed as DC amplifier and connected with the output stage of the transmitter in the manner shown in *fig. 13.* In this way all the sources of supply (also those of the filaments) of the transmitter output stage are at the positive anode voltage (about 1500 V) of the modulator with respect to earth, which necessarily leads to a somewhat unusual construction of the supply installation.

As regards the final application of the DC voltage component in the modulator stage, that component can be transmitted through a separate cable from the studio to the transmitter, as is indeed done in certain transmitters. In the case of the transmitter in Eindhoven, however, a much simpler but nevertheless very satisfactory method is applied. In *fig. 14,* the variation of the grid voltage v_g of the modulator stage, as it is desired is represented for a completely white, a completely black and any given picture. The differences in the DC voltage component with the different types of picture may be seen. If we apply this signal without the DC voltage component to the modulator stage with the help of the normal grid condenser, while the modulator stage is given a fixed control grid bias v_{g_0}, the grid voltage for a completely white and a completely black picture would have the form drawn in *fig. 15.* Instead of this, however, we can also obtain the desired form according to fig. 14 if instead of giving the modulator a fixed grid bias, we give it a bias which can be derived from the incoming signal itself with the help of the grid condenser C_1 and a leakage

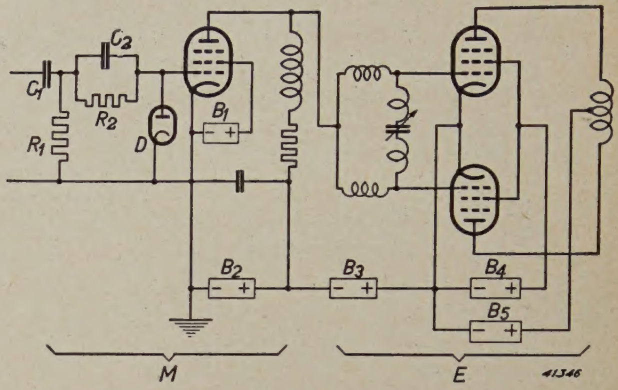

Fig. 13. The modulator stage (M) is constructed as a DC amplifier in order to be able to apply the DC voltage component of the video-frequency signal to the grid of the transmitter output stage (E). As a result all the sources of supply of the output stage are at the positive anode voltage of the modulator with respect to earth.
B_1 screen-grid supply, modulator, B_2 anode supply, modulator, B_3 source of grid bias, output stage, B_4 screen grid supply, output stage, B_5 anode supply, output stage.

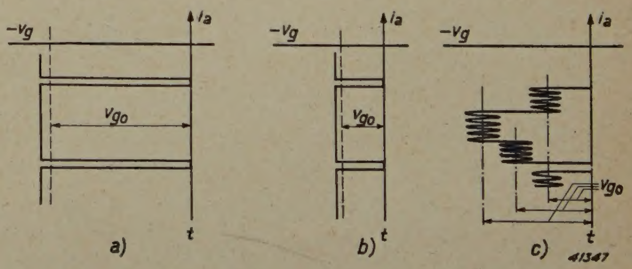

Fig. 14. The desired form of the signal on the grid of the modulator stage, a) for a completely white picture, b) for a completely black picture, c) for any given picture. The differences in the DC voltage component (v_{g0}) may be seen, and in case (c) they may also vary within a picture. (The line synchronization signals are again omitted.)

resistance R_1. The bias then adjusts itself automatically so that the voltage peaks reach just to the region of the grid currents ($v_{g0} \approx 0$). Care must be taken however, that the bias thus obtained can be altered rapidly enough to follow any variations of the average brightness in the picture. "Rapidly enough" means that the picture is divided, for instance, into four strips parallel to the direction of the lines and the average brightness is correctly reproduced for each of these strips. Now the possible velocity of variation of the bias is higher, the smaller the product R_1C_1. This product may not, however, become too small, since otherwise the lowest modulation frequencies are not faithfully transmitted, for a lowest modulation frequency of 25 c/sec R_1C_1 must be equal to 0.4, while in order to follow the fluctuations of the average brightness a value at least 20 times as small is required. This difficulty could be solved by introducing a second circuit after C_1R_1 in the manner indicated in fig. 13, with a small C_2 and a large R_2 ($R_2 = 100\ R_1$) of which the product R_2C_2 amounts to only about 0.01. This circuit has practically no effect on the transmission of the video-frequencies; the grid bias, however, can now be considered to consist of the practically constant contribution of the average voltage over R_1 and the contribution of the sufficiently rapidly varying average voltage over R_2. The latter assumes the correct value at every moment due to the fact that at the voltage peaks grid current flows for a moment. Since the adjustment will be more exact, according as the current in the grid circuit grows more rapidly upon the approach of the grid voltage to zero (it already begins to flow at a certain negative grid voltage), a small diode is connected between grid and cathode as indicated in fig. 13, which diode begins to pass current as soon as its anode voltage has nearly become positive.

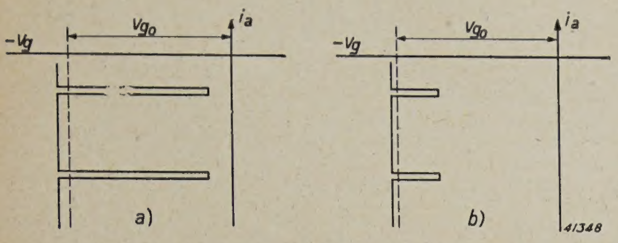

Fig. 15. Signal on the grid of the modulator stage, as it would be obtained upon suppression of the DC voltage ocmponent and with a fixed negative grid bias *a*) for a completely white, *b*) for a completely black, picture.

THE INFLUENCE OF LOSSES ON THE PROPERTIES OF ELECTRICAL NETWORKS

621.318.7

by J. F. SCHOUTEN and J. W. KLÜTE.

Two methods are discussed for determining the influence of losses on the properties of electrical networks such as filters. Especially the method of stretching membranes between certain space curves furnishes a very graphic picture of the attenuation and the variation of the phase rotation of a network as a function of the frequency and the measure of loss.

In calculating electrical networks, and especially electrical filters, it is usually assumed that the self-inductions and capacities used in them are ideal. By ideal is meant that the impedances or admittances of these elements are purely imaginary, so that no losses occur. For these ideal elements the properties of a filter can then be calculated according to known methods [1]).

In practice, however, we are always concerned with elements which are subject to losses, which can be represented for instance by a resistance in series with the ideal coil or a resistance in parallel with the ideal condenser. The properties of a filter which is built up of these "actual" elements can immediately be expressed in mathematical form [2]). The numerical interpretation of this form is ,however, very elaborate and time-consuming.

In this article we shall discuss two methods by which a general and in part very graphic impression is given of the changes undergone by the properties of a filter calculated for ideal elements when the influence of the losses is taken into account [3]).

The method of the perpendicular derivative

The transition from the ideal to the actual self-induction amounts to the fact that the impedance Z, instead of being purely imaginary, becomes complex (see *fig. 1*), which may be written as follows:

$$Z = j\omega L \rightarrow R + j\omega L, \quad . \quad . \quad . \quad . \quad . \quad (1a)$$

where R represents the series resistance of the coil.

In the same way, for the admittance Y of the condenser:

$$Y = j\omega C \rightarrow G + j\omega C, \quad . \quad . \quad . \quad . \quad (1b)$$

where G represents the admittance of the parallel resistance of the condenser. In general the quantities L, R, C and G will depend more or less upon the frequency, but we shall not go into that here.

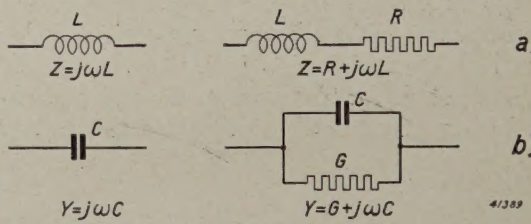

Fig. 1. Equivalent diagram of the ideal (left) and actual (right) elements.
a) The actual self-induction may be considered as an ideal self-induction L in series with a real impedance R.
b) The actual capacity may be considered as an ideal capacity C in parallel with a real admittance G.

We now write formulae (1a) and (1b) in the following form :

$$\left.\begin{aligned} Z &= R + j\omega L = \left(\frac{R}{L} + j\omega\right) L = (\varrho + j\omega)L, \\ Y &= G + j\omega C = \left(\frac{G}{C} + j\omega\right) C = (\gamma + j\omega)C. \end{aligned}\right\} \quad (2)$$

In other words, the transition from the ideal to the actual coil, retaining the self-induction L, amounts only to the substitution of $\varrho + j\omega$ for the quantity $j\omega$. In the former $\varrho = R/L$ represents the measure of loss of the coil. In the same way the transition from the ideal to the actual condenser, retaining the capacity C, amounts only to the substitution of the quantity $\gamma + j\omega$ for $j\omega$. Now $\gamma = G/C$ represents the measure of loss of the condenser. Let us for the present assume that the measure of loss of the coils and condensers in the network is the same and equal to k. The following conclusion may then be drawn. If the properties of an ideal network are known as functions of the frequency and of the values of the various self-inductions and capacities, the properties of the actual network follow from them directly by replacing $j\omega$ everywhere in the formulae by the complex quantity $k + j\omega$. Thus if a property of the ideal network is represented by an analytic function of the argument $j\omega$, which we shall call $f(j\omega)$, the same property for the actual network is represented by the same function of the complex argument $k + j\omega$, thus by $f(k + j\omega)$.

---

[1]) Balth. van der Pol and Th. J. Weijers, Electrical filters, Philips Techn. Rev. **1**, 240, 270, 298 and 327, 1936. Referred to below as F.

[2]) Loc. cit. F, p. 302.

[3]) These considerations are generally valid not only for electrical networks, but also for other systems, mechanical for instance. Here, however, we shall confine ourselves in terminology and examples to electrical filters.

The function $f(k + j\omega)$ is in general complex. It may therefore be written as the sum of a real part α and an imaginary part β:

$$f(k + j\omega) = \alpha + j\beta.$$

For such a complex function the following is valid, and may easily be verified by differentiation with respect to k and ω, respectively:

$$\left.\begin{aligned} \frac{\delta\alpha}{\delta k} &= \frac{\delta\beta}{\delta\omega} \\ \frac{\delta\beta}{\delta k} &= -\frac{\delta\alpha}{\delta\omega} \end{aligned}\right\} \quad \ldots\ldots\ldots \quad (3)$$

These are the well-known differential equations of Cauchy-Riemann from the theory of the functions of a complex variable. These equations may now be interpreted as follows.

The properties of our network, expressed as the sum of a real part α and an imaginary part β, are known as a function of the frequency ω for the case when $k = 0$ (the ideal network). We wish ,however, to know the properties for a value of the measure of loss k which differs somewhat from zero (the actual network). The equations (3) show that for a given frequency ω the required change of α with k is equal to the known change of β with ω for the same frequency. In the same way the required change of β with k is equal to the known change of α with ω for the same frequency but with the opposite sign. If we imagine two systems of rectilinear coordinates in which the quantities α and β, respectively, are plotted as functions of k and ω, relation (3) means that at corresponding points (k, ω) the derivative in the direction k on the one surface is equal (with the positive or negative sign) to the derivative in the direction ω perpendicular to the first on the other surface.

If we develop α and β at a certain frequency ω in a Taylor's series in powers of k and apply relations (3), we find as a first approximation [4]:

$$\left.\begin{aligned} \alpha(k) &= \alpha(0) + k\frac{\delta\alpha(0)}{\delta k} = \alpha(0) + k\frac{\delta\beta(0)}{\delta\omega}, \\ \beta(k) &= \beta(0) + k\frac{\delta\beta(0)}{\delta k} = \beta(0) - k\frac{\delta\alpha(0)}{\delta\omega}. \end{aligned}\right\} \quad (4)$$

With this, for not too large values of k, we can determine $\alpha(k)$ as well as $\beta(k)$ from the given dependence of $\alpha(0)$ and $\beta(0)$ on the frequency ω.

The choice of α and β

We have expressly left open the question of which function we wish to choose to characterize our network. We may therefore still make a free choice. In the case of an impedance we may, for example, take for α and β the real and the imaginary parts of the impedance itself, but equally well those of the logarithm of that impedance, or, if it were desirable, the real and the imaginary parts of the square or the e-power of the impedance. For all these cases relations (3) are valid.

It is customary to characterize electrical filters (quadripoles) as follows [5]. An input voltage $e^{j\omega t}$ gives rise to the following output voltage :

$$e^{-T}\, e^{j\omega t} = e^{-\alpha - j\beta + j\omega t} \quad \ldots \quad (5)$$

In this α represents the attenuation, *i.e.* the logarithm of the ratio of the absolute values of input and output voltage, while β indicates the phase rotation. For this α and β (whose nomenclature was anticipated in our choice of symbols), which represent the real and the imaginary part of the propagation constant T, equations (3) are thus valid.

Examples

We shall now explain relations (3) and (4) by means of several examples.

a) The low-pass filter of the basic type [6]

The ideal low-pass filter of the basic type (*fig. 2a*) has the following properties. The attenuation α is equal to zero below the limiting frequency ω_1. Below this limit therefore all frequencies are passed without attenuation (transmission region). Above the limiting frequency (attenuation region) the attenuation rises rapidly (fig. 2*b*) according to the following relation:

---

[4] The relation (4) was first given by H. F. Mayer E.N.T. **2**, 335, 1925. To E. A. Guillemin, Communication Networks II, p. 445 is due the remark that they are ultimately based upon the theory of complex functions.

[5] Loc. cit. F, p. 270.

[6] Loc. cit. F, p. 298.

$$\alpha = 2 \cosh^{-1} \frac{\omega}{\omega_1}.$$

The phase β in the transmission region is given by

$$\beta = 2 \sin^{-1} \frac{\omega}{\omega_1}.$$

At $\omega = \omega_1$, β therefore attains the value π; this value is retained for the whole damping region (fig. 2*c*). From the given behaviour of β as a function of ω we may draw the following conclusions about the influence of the measure of loss k on the attenuation α:

1) In the damping region the attenuation α is in the first approximation independent of the measure of loss k, since $\delta\beta/\delta\omega = 0$.
2) In the low part of the transmission region ($\omega < 2/3\ \omega_1$) the attenuation α is independent of the frequency, directly proportional to the measure of loss k (eq. 4) and inversely proportional to the width ω_1 of the filter, since $\delta\beta/\delta\omega \approx 2/\omega_1$ (fig. 2*c*).
3) In the high part of the transmission region ($\omega < 2/3\ \omega_1$) the increase in the attenuation is greater than in the low part, considering the increase of $\delta\beta/\delta\omega$ with ω.

For the influence of the measure of loss k on the phase β we find in the same way:

1) In the transmission region the phase β is in a first approximation independent of the measure of loss k, since $\delta\alpha/\delta\omega = 0$.
2) In the damping region the phase β decreases, since $\delta\alpha/\delta\omega$ is positive and directly proportional to the measure of loss k. This decrease becomes smaller with increasing frequency, since $\delta\alpha/\delta\omega$ decreases.

Before passing on to a numerical example it must be pointed out that our original condition, namely that the measures of loss of coils and condensers are equal, is practically never fulfilled in practice. On the contrary the losses of the condensers are usually many times as small as those of the coils. Our formulae (3) and (4) then, nevertheless, remain valid in the first approximation provided the average of ϱ and γ is taken for k, thus:

$$k = \frac{\varrho + \gamma}{2}, \text{ so that } k = \frac{\varrho}{2}, \text{ when } \varrho \gg \gamma.$$

The influence of the inequality of ϱ and γ, which is ignored here, is manifested mainly in a slight shift of the frequency.

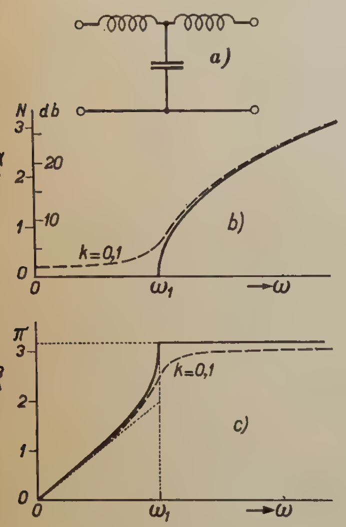

Fig. 2.

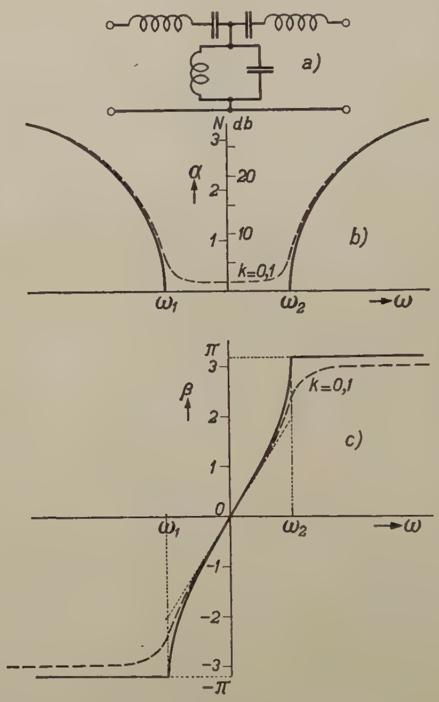

Fig. 3.

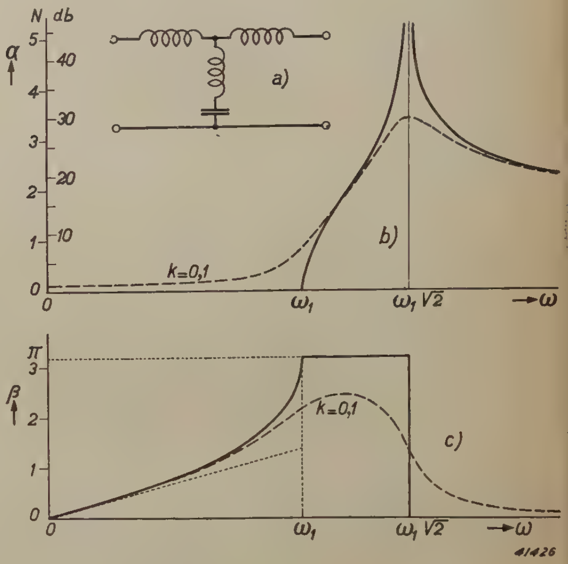

Fig. 4.

Figs. 2, 3 and 4. Low-pass and band-pass filter of the basic type and low-pass filter of the transformation type ($\omega_\infty/\omega_1 = \sqrt{2}$).
a) *T*-cells of the filter types in question.
b) and *c*) Behaviour of the attenuation α and phase β, respectively, as functions of the frequency for the ideal case when $k = 0$ (full-drawn) and for $k = 0.1\ \omega_1$ (broken line).

We now take as example a low-pass filter consisting of a number of cells. We assume a cut-off frequency of $\nu_1 = 400$ c/sec, thus $\omega_1 = 25\,000$, a measure of loss of the coils of $\varrho = R/L = 125$ ohms per henry and a negligable measure of loss of the condensers. In the low part of the transmission region, independent of the frequency, the damping becomes

$$\alpha = k\,\frac{\delta\beta}{\delta\omega} = \frac{\varrho}{2}\cdot\frac{2}{\omega_1} = \frac{1}{200}\ \text{neper/cell.}$$

The fact that the attenuation here is found in neper per cell is a result of the fact that we began (see (5)) with the natural logarithm. If we wish to express the damping in decibels, since 1 neper corresponds to 8.7 decibels, we find that $\alpha = 0.043$ db/cell. The same formulae are valid for a loaded cable consisting of a number of sections, which may of course also be considered as a low-pass filter.

b) *The band-pass filter of the basic type* [7]

For this filter practically the same considerations hold as for the low-pass filter. For the damping in the middle section of the transmission region we now find (*fig. 3*):

$$\alpha = 8.7\,k\,\frac{\delta\beta}{\delta\omega} = 8.7\,\frac{4k}{\omega_2 - \omega_1}\ \text{db/cell} \quad . . \; (6)$$

From this we draw the conclusion that the attenuation in the transmission region of the band-pass filter is independent of the mean frequency of the transmitting band and that it is determined only by the measure of loss k, to which the attenuation is directly proportional, and by the width $\omega_2 - \omega_1$ of the band, to which it is inversely proportional.

In connection with this it may be pointed out that the losses of a coil are often expressed not in the terms of the measure of loss ϱ but in those of the quality Q[8]), which is related as follows to the measure of loss:

$$Q = \frac{\omega L}{R} = \frac{\omega}{\varrho},\ \text{or}\ \varrho = \frac{\omega}{Q}.$$

If one now considers the case often occuring in carrier-wave telephony, namely a number of band-pass filters at different points but with the same band width, these filters undergo the same change in α and β when the values of ϱ and not those of Q are the same for the different transmission bands. Considered from the point of view of filter theory, therefore, it is better to use ϱ and not Q to characterize the losses. For the same reason we prefer the measure of loss γ for condensers over the customary tangent[8]) of the phase displacement δ, which is related to γ as follows:

$$\tan\delta = \frac{\omega C}{G} = \frac{\omega}{\gamma},\ \text{or}\ \gamma = \frac{\omega}{\tan\delta}.$$

c) *The low-pass filter of the transformation type* [9])

Compared with the low-pass filter of the basic type, this filter has the peculiarity that the attenuation already becomes infinite for a finite frequency ω_∞ (in *fig. 4* $\omega_\infty = \omega_1\sqrt{2}$) and above this frequency gradually falls to a certain limit. It is used when a very steep rise in the damping is required immediately above the cut-off frequency ω_1. In the phase β the frequency of infinite attenuation is manifested in that the phase jumps back abruptly from π to zero. This means, therefore, that in the whole damping region, as for the filter of the basic type, $\delta\beta/\delta\omega$ is zero except at ω_∞, where this quantity becomes negatively infinite. This makes it understandable that here also in the damping region the damping α is in the first approximation everywhere independent of the measure of loss k except exactly at point ω_∞, where the smallest loss already causes the damping to decrease to a finite value.

In the transmission region the phase does indeed vary from zero to π, but more slowly at first, and consequently more rapidly later, than is the case with the basic type. This means that in the low part of the transmission region this type of low-pass filter is less sensitive and in the high part more sensitive to losses than the basic type (fig. 4*b*). This is even truer the closer the point of infinite damping ω_∞ lies to the cut-off frequency ω_1. This influence of the point of infinite damping on the behaviour of the filter in the transmission region must here be accepted as given. In the following, however, on the treatment of the problem with the help of membrane models we shall reach a closer insight into this point.

With the influence of the losses on the phase β we must also note that for frequencies above ω_∞ $\delta\alpha/\delta\omega$ is negative. In this region therefore β must increase with increasing measure of loss k.

These considerations illustrate how with the help of the method of the perpendicular derivative a fairly complete concept may be formed of the behaviour of electrical systems which are subject to slight losses, when the behaviour of the system free of losses is known. We shall now derive another relation between the quantities α, β, k and ω which will lead us to a direct graphic representation of α and β.

) Loc. cit. F, p. 327.

[8]) Instead of the quantities Q and $\tan\delta$ the quantities d_L and d_C (loc. cit. F. p. 302), respectively, are also used. These latter quantities are related as follows:

$$d_L = 1/Q \text{ and } d_C = \tan\delta.$$

[9]) Loc. cit. F. Table opposite p. 332.

The method of the membrane model

By differentiation of equations (3) with respect to k and ω we can derive the following relations:

$$\left.\begin{aligned}\frac{\delta^2\alpha}{\delta k^2}+\frac{\delta^2\alpha}{\delta\omega^2}&=0,\\ \frac{\delta^2\beta}{\delta k^2}+\frac{\delta^2\beta}{\delta\omega^2}&=0.\end{aligned}\right\}\quad\ldots\ldots\ (7)$$

These are Laplace differential equations, also well known from the theory of the functions of a complex variable.

These equations do not, like equations (3), represent a relation between α and β, but a condition which not only α but also β, each as a function of k and ω, must fulfil. In spatial representation therefore they represent conditions which are applied to the surfaces $\alpha(k,\omega)$ and $\beta(k,\omega)$. The Laplace equation exhibits great similarity with the differential equation which determines the shape of a soap film, that is to say of the so-called minimum surface. The name minimum surface comes from the fact that such a surface is the smallest which can be laid through a given contour. For very small slopes the differential equation for the minimum surface passes over into that of Laplace. Use has already been made of this in this laboratory, namely for the construction of a model of a two-dimensional potential field by means of a rubber membrane [10]). The behaviour of the rubber membrane is slightly different from that of the soap film, since the tension at every point of the membrane, unlike that of the soap film, is not the same, but depends upon the stretch, which differs from point to point. For small slopes, however, it also satisfies the Laplace equations.

For the surfaces α and β about to be considered the condition that the slopes must be small is by no means satisfied. We shall, however, for the present disregard the inaccuracies thereby caused, and we may then, although roughly, reproduce the form of the surfaces $\alpha(k,\omega)$ and $\beta(k,\omega)$ by representing them as a stretched soap film or rubber membrane.

Just as the solution of the Laplace equation in the interior of a closed contour is fully determined by the values of the function at the edge, in the same way the form of a soap film or of a rubber membrane is completely fixed by its contour. In the practical execution therefore we shall try to determine α and β along a contour, then we shall make a model of it and over it we shall stretch a soap film or a rubber membrane.

Examples

The method of the membrane model will now be explained with the aid of several examples.

a) The low-pass filter of the basic type

We have already stated that for the ideal filter (thus in our spatial representation in the coordinates α, k and ω for the plane $k = 0$) the following is valid:

$$\alpha = 2\cosh^{-1}\frac{\omega}{\omega_1}\quad\ldots\ldots\ (8)$$

Similarly we find for the plane $\omega = 0$

$$\alpha = 2\sinh^{-1}\frac{k}{\omega_,}\quad\ldots\ldots\ (9)$$

and for a plane of constant value α

$$\frac{\omega^2}{\omega_1^2\cosh^2\frac{\alpha}{2}}+\frac{k^2}{\omega_1^2\sinh^2\frac{\alpha}{2}}=1.\ \ .\ \ (10)$$

In *fig. 5* a model is shown of the surface $\alpha(k,\omega)$ where the rubber membrane is stretched over the given contours (8), (9) and (10). Here we again encounter the complete behaviour which has already been described in the method of the perpendicular derivative. In the

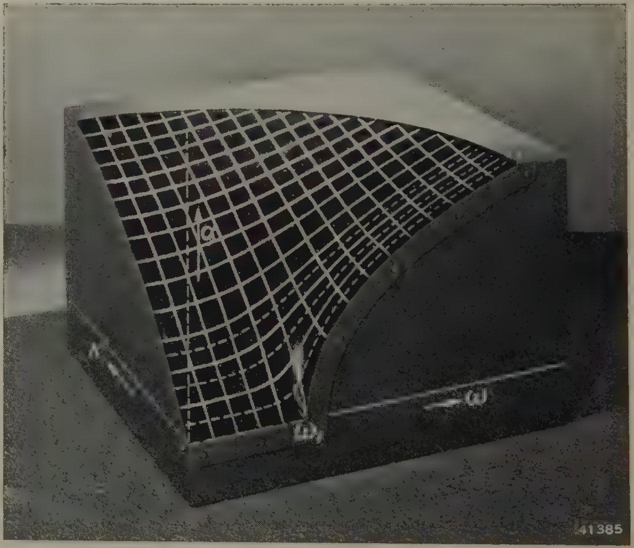

Fig. 5. Rubber model of the damping α (k,ω) for the low-pass filter of the basic type. It should be noted how the attenuation region on the axis $k = 0$ draws the membrane upward for the high part of the transmission region and for $k \neq 0$.

---

[10]) P. H. J. A. Kleynen, Philips Techn. Rev. **2**, 338, 1937.

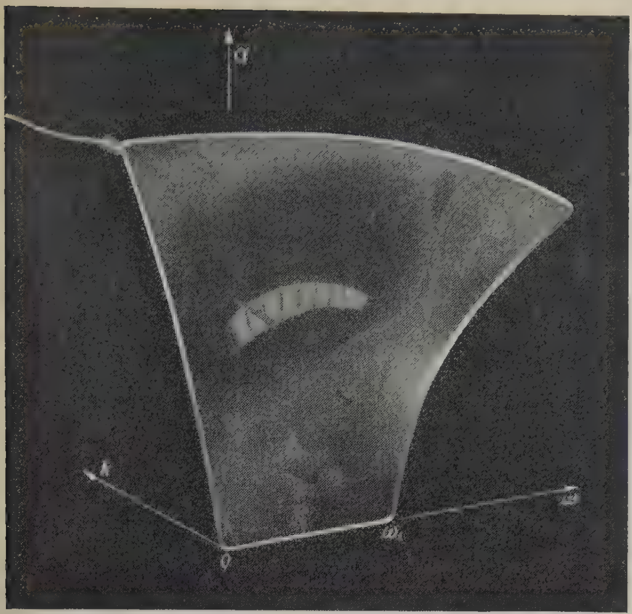

Fig. 6. Soap-film model of the damping α (k, ω) for the low-pass filter of the basic type.

model the values for constant k and constant ω are indicated by white lines. It may clearly be seen how in the low part of the transmission region α increases linearly with k. As, however, we approach the cut-off frequency ω_1, the neighbouring attenuation region already begins to draw the membrane more upwards, so that at $\omega = \omega_1$ the derivative $\delta\alpha/\delta k$ even becomes infinite. Once in the damping region the derivative immediately falls to zero, the membrane is here exactly perpendicular to the plane $k = 0$. In *fig. 6* the same surface is represented in the form of a soap film.

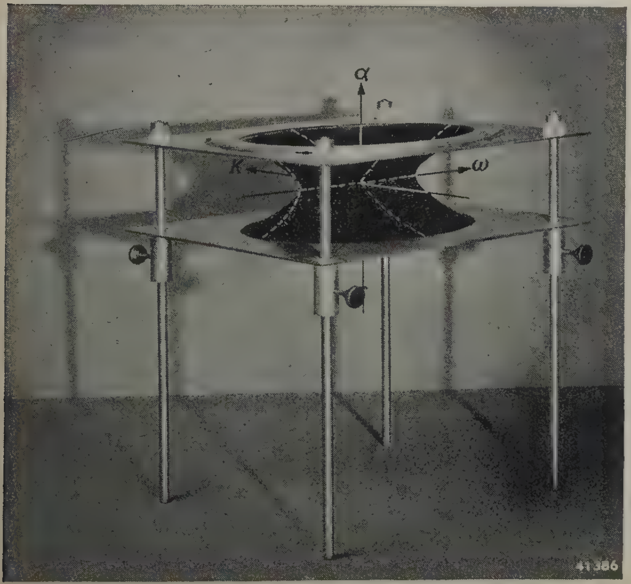

Fig. 7. Rubber model of the complete surface α (k, ω) for the low-pass filter of the basic type (extended to negative valvues of α, k and ω). The model consists of two rubber membranes stretched over two ellipses and fastened to each other along the line which connects the two foci of each ellipse. In the plane $k = 0$ and $\omega = 0$ the surface must automatically assume the form of a hyperbolic cosine and hyperbolic sine respectively. This condition is not completely satisfied by the rubber model, but with a certain approximation.

We may also imagine the surface to be continued for negative values of α, k and ω. Lines of constant α are then according to (10) given by a series of confocal ellipses, *i.e.* ellipses with the common foci $-\omega_1$ and $+\omega_1$. We now cover two similar ellipses with a rubber membrane and place the ellipses vertically above each other in a horizontal position (*fig. 7*). If we now fasten the two membranes to each other along the line which joins the foci of each ellipse we shall obtain the desired surface. The variation according to a hyperbolic cosine in the plane $k = 0$ and according to a hyperbolic sine in the plane $\omega = 0$ must then automatically be attained.

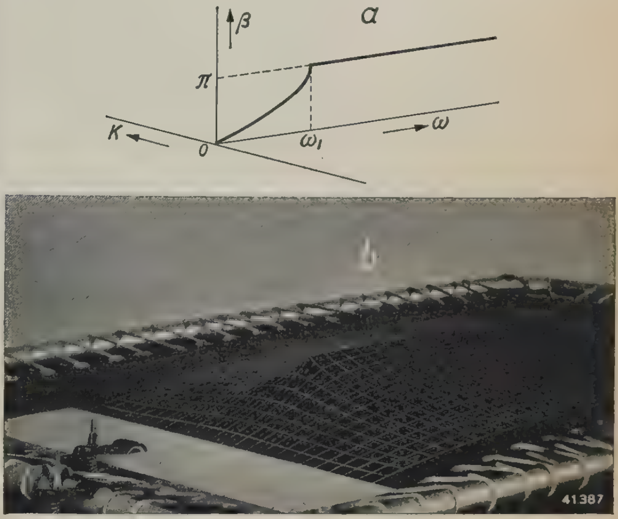

Fig. 8. Rubber model of the phase β (k, ω) for the low-pass filter of the basic type for the values of β between 0 and π.
a) Diagram of the surface β (k, ω).
b) The rubber model of the phase β.
The membrane is fixed by three conditions: the value $\beta = 0$ along the periphery, the value $\beta = 0$ along the axis $\omega = 0$ (fixed by a straight strip above the surface) and the value $\beta = \pi$ along the axis $k = 0$ for $\omega > \omega_1$ (fixed by a straight strip under the surface). The variation according to an arc sine in the transmission region must then be automatically attained.

For the phase β we can construct a similar model. We may begin by determining that for the frequency 0 the phase must always be zero for each k, since the phase β is always an antisymmetric function of ω (*i.e.* a function which for negative values of the argument takes on a value equal and opposite to that for positive values of the argument). Furthermore it can be proved that, for a large value of k, β approaches

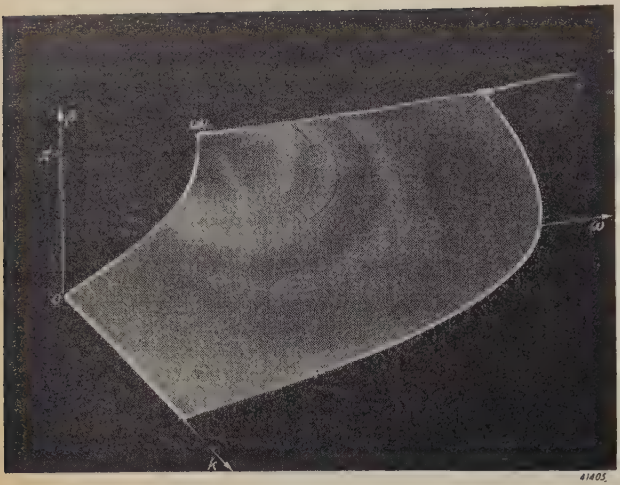

Fig. 9. Soap-film model of the phase β (k, ω) for the low-pass filter of the basic type. The coordinate k is here chosen in the opposite direction for the sake of clearness.

zero independent of ω. Sufficient data are thus available to fix completely the surface β. For this purpose we take a rubber membrane (*fig. 8*) whose fixed periphery approximately represents the phase zero. If now by means of a vertical straight strip under the membrane we push up the latter along the ω-axis above the cut-off frequency ω_1 to a certain height which represents the height π, and by means of a second strip above the membrane we keep it down along the k-axis at the height zero, the surface β is entirely fixed. The rectangular network of white lines drawn on the membrane now represents the set of lines of constant k and constant ω respectively. The variation according to an arc sine on the line $k = 0$ for the region between the origin and the cut-off frequency must be automatically assumed.

In *fig. 9* the same function is represented as a soap-film model, but only for positive values of k.

b) The band-pass filter of the basic type

The band-pass filter of the basic type, when the width of the transmission region is small compared with the average frequency of the transmission region, is entirely analogous to the low-pass filter of the same type. It may be considered as a low-pass filter which is reproduced not only for positive but also for negative values of the frequency, and which is shifted in such a way toward high frequencies that the frequency zero is displaced to the mean frequency ω_0 of the transmission region. The limits $-\omega_1$ and $+\omega_1$ of the low-pass filter are then replaced by the limits ω_1 and ω_2 of the band-pass filter.

In *figs. 10* and *11* we show drawings of models of the surfaces α and β. In *fig. 12* the surface α is shown as a soap film.

c) The low and band-pass filter of the transformation type

There is the same relation between these

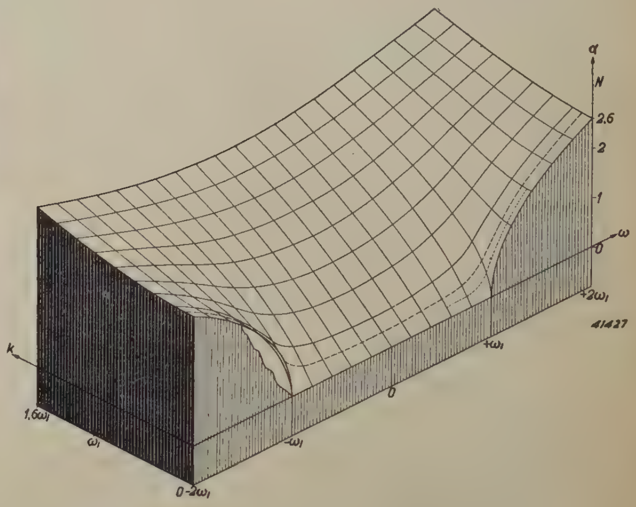

Fig. 10.

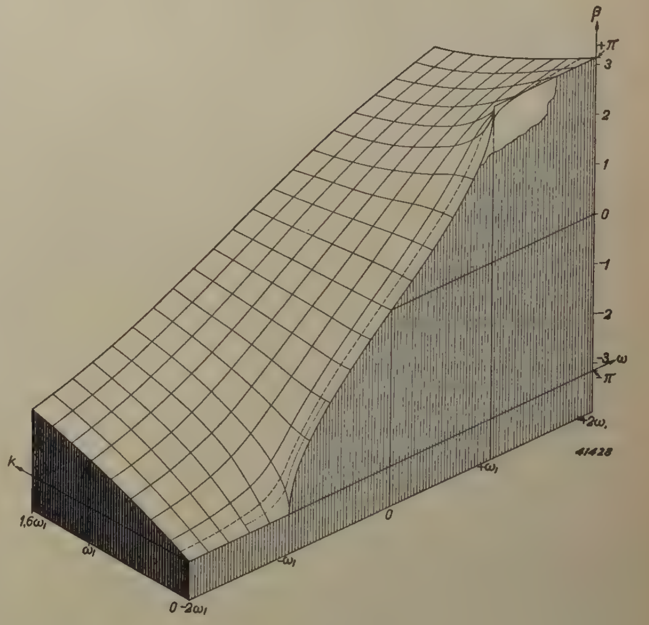

Fig. 11.

Figs. 10 and 11. Drawn models of the attenuation α (k, ω) and the phase β (k, ω) of the low-pass filter of the basic type [11]). By approximation the models also hold for a narrow band-pass filter of the basic type when the mean frequency ω_0 of the transmission region is added to each frequency. The values for $k = 0.1\ \omega_1$ are indicated by a dash line, those for $k = 0.05\ \omega_1$ by a dotted line.

---

[11]) The models are exact representations of the real and the imaginary part of the function

$$T = \alpha + j\beta = 2 \sinh^{-1}\left(\frac{k + j\omega}{\omega_1}\right).$$

two filters as between the corresponding filters of the basic type. In *figs. 13* and *14* drawings of models of the surfaces α and β are reproduced. In *fig. 15* the surface α is represented as a soap film. Our feeling for the behaviour of the soap film tells us immediately that for $k \neq 0$ the film draws the attenuation α in the transmission region upwards, but in the vicinity of the points of infinite attenuation it draws it downward. A fundamental difference from the band-pass filter of the basic type consists in the fact that for high values of k the attenuation α does not tend towards an infinitely high value, but towards a finite value independent of ω. These low values at the edge of the film will now have a tendency to cause the film to run upwards less strongly in the transmission region than is the case with the basic type. In the vicinity of the cut-off frequencies, however, this influence is overcome by the nearness of the points of infinite damping, which draw the film more strongly upward than is the case with the basic type. This results finally in the smaller influence of the losses in the middle and the stronger influence at the edges of the transmission region of the filter. We have already pointed this out when discussing the low-pass filter of the transformation type according to the method of the perpendicular derivative. In *fig. 16* a soap-film model is shown of the phase of the low-pass filter, while in *fig. 17* a rubber model of the same filter is shown. In the latter model in addition the second part of the surface β is shown which extends from π to 2π. The model is constructed in a way analogous to that of fig. 7. The two membranes are fastened together along the line from ω_1 to ω_∞, and a vertical plate is then inserted between the two surfaces

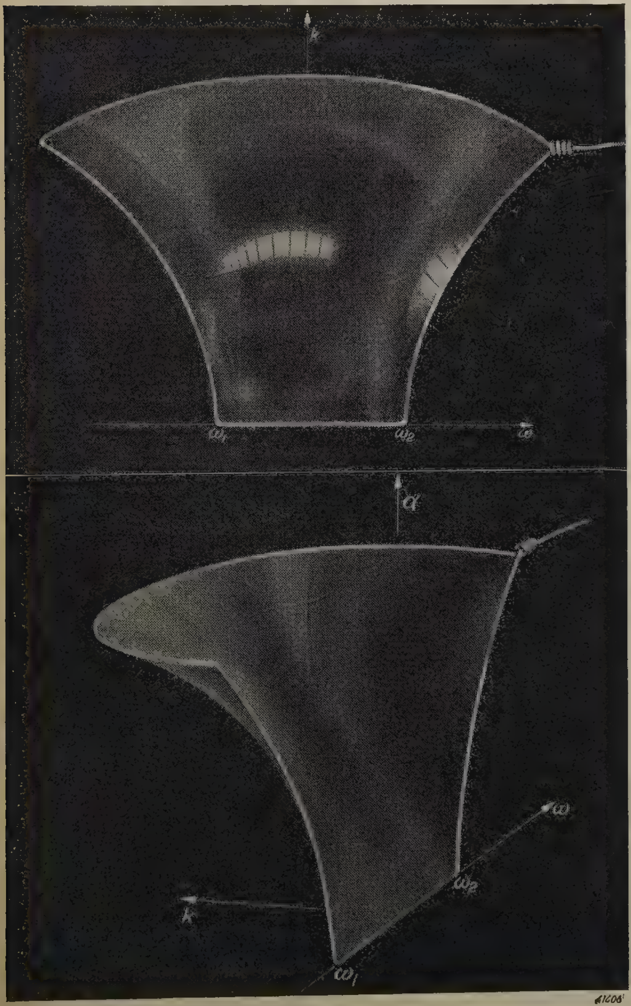

Fig. 12. Soap-film model of the damping α (k, ω) of the band-pass filter of the basic type.

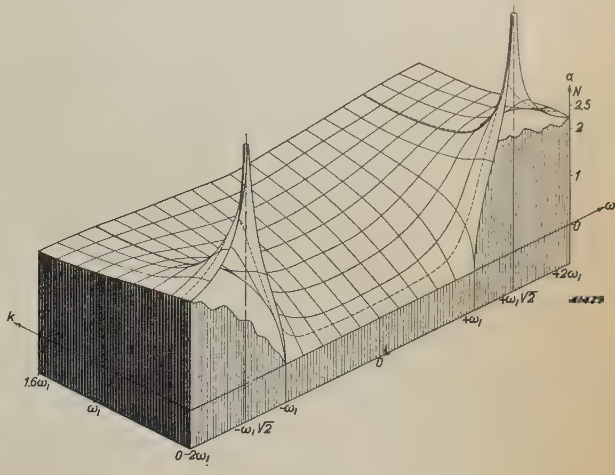

Fig. 13.

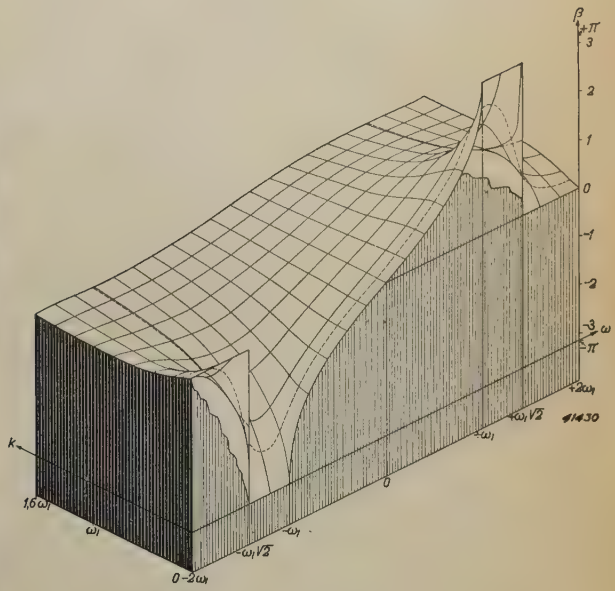

Fig. 14.

Figs. 13 and 14. Drawn models of the damping α (k, ω) and the phase β (k, ω) of the low-pass filter of the transformation type $(\omega_\infty = \omega_1\sqrt{2})$. The models also hold for the band-pass filter of the transformation type in the manner described under figs. 10 and 11.

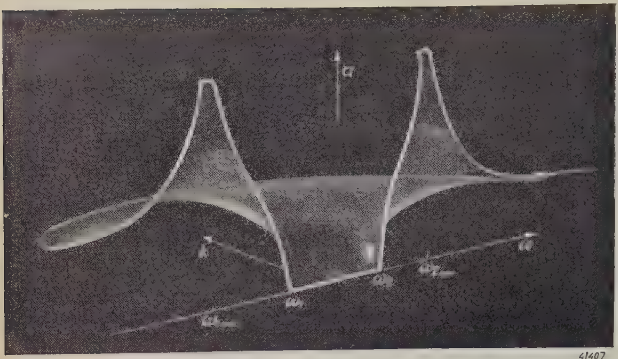

Fig. 15. Soap-film model of the attenuation α (k, ω) for the band-pass filter of the transformation type.

in order to keep the value of β on the ω-axis for frequencies above ω_∞ at zero for the lower and at 2π for the upper surface.

d) A compound band-pass filter for carrier-wave telephony

In carrier-wave telephony use is made of band-pass filters whose chief function is to suppress one of the two sidebands formed during modulation. For this purpose a damping characteristic is chosen which rises steeply on one side and less steeply on the other. For such an asymmetric filter [12]), which has recently been described in this periodical and which consists of three different cells, a soap-film model is shown in *fig. 18*. It is clear that the influence of the losses at the steep edge of the transmission region is greater than at the less steep edge.

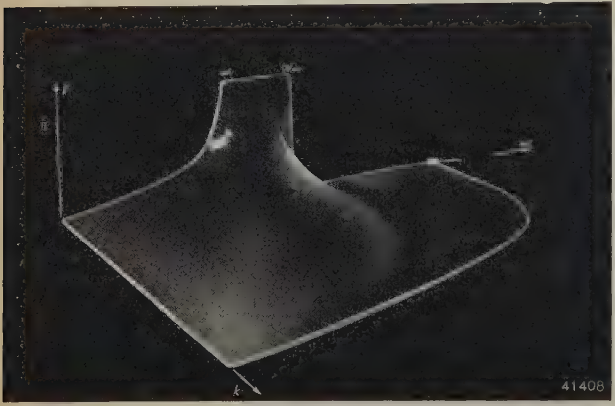

Fig. 16. Soap-film model of the phase β (k, ω) for the low-pass filter of the transformation type. The coordinate k, as in fig. 9, is here chosen in the forward direction.

The relation between α and β

Upon closer consideration of the results which were produced by the method of the perpendicular derivative and that of the film model, we encounter a remarkable situation. In the first method we deduced for instance that $\delta\alpha/\delta k$ is given by $\delta\beta/\delta\omega$. In the second method, however, we gave α along a contour only and from it derived α unambiguously in the whole inner region, thus also $\delta\alpha/\delta k$ along the axis $k = 0$. Thus the given contour in fig. 5 furnished

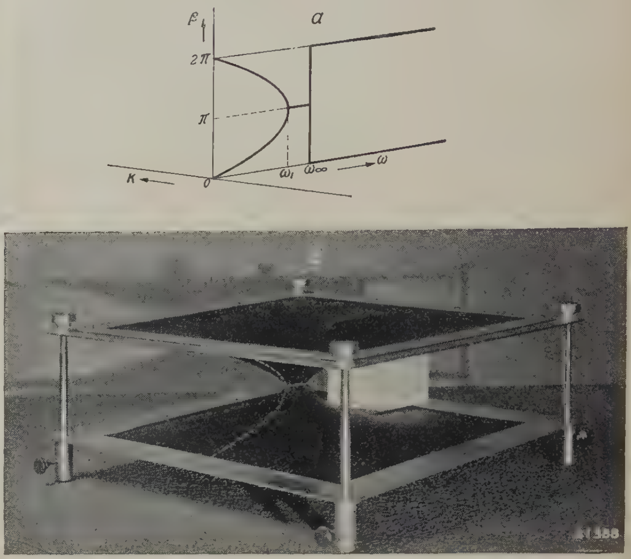

Fig. 17. Rubber model of the phase β of the band-pass filter of the transformation type for values of β between 0 and 2π.

a) Diagram of the surface β.

b) Rubber model of the phase β for the region between 0 and 2π.

The two sheets of the surface are fixed by three known conditions: the values zero and 2π, respectively, along the edges (to which the axis $k = 0$ also belongs), the value π on the ω-axis between ω_1 and ω_∞ (obtained by fastening the two sheets together) and the values zero and 2π, respectively, along the ω-axis for frequencies higher than ω_∞ (obtained by the vertical plate, which, in order to prevent undue damage to the membranes, is not inserted quite as far as the value ω_∞).

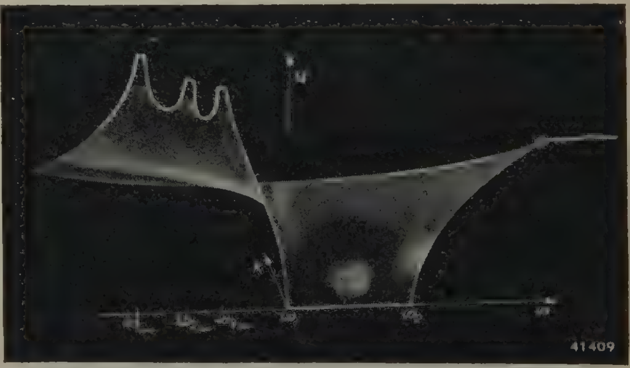

Fig. 18. Soap-film model of the attenuation α (k, ω) of a composite band-pass filter for carrier-wave telephony.

---

[12]) Th. J. Weijers, Philips Techn. Rev. 7, 104, 1942.

us with the surface α for the low-pass filter without it being necessary to use our knowledge of the behaviour of β as a function of ω. That this is possible is due to the fact that α and β are not independent quantities. With a given damping as a function of the frequency, the variation of the phase cannot be chosen freely. The explanation of this relation, which we shall not go into here, ultimately comes down to the simple fact that whatever the properties of the filter it is always known with certainty that no output voltage can occur before an input voltage is applied. From this the relation between α and β may be derived.

GENERATORS FOR SHORT-WAVE THERAPY

by J. FRANSEN and J. M. LEDEBOER.

615.846

In the medical treatment with short waves (short-wave therapy) use is made of the fact that by means of high-frequency currents with frequencies above 10^7 c/sec a development of heat can be obtained at any desired spot in the body of the patient. The high-frequency currents are excited by means of short-wave transmitters which in some instances are simpler and in others more complicated than the transmitters used for radio and television. The requirements made of such transmitters and their construction are discussed here on the basis of a description of two generators for short-wave therapy developed by Philips. These generators work on a wavelength of 6 m and can deliver a maximum effective output of 400 and 150 W, respectively, to the patient.

Introduction

Around about 1800, when hardly anything was yet known about the great field of electrical phenomena, attempts were already being made to send electric currents through the human body for the purpose of producing a healing effect upon the sick. Direct current (galvanism) was originally used, later on alternating current of 20 to 200 c/sec (faradism). A specific effect of the electric current was indeed found, namely a prickling sensation which sometimes in the case of nervous disorders was found to have therapeutic effect. At the same time, however, electrolysis occurs in the tissue, and this was the main reason why the current could not be made stronger than 30 or 40 mA.

An increase of the current used only became possible when in 1890 d'Arsonval discovred that with alternating current of a high frequency (above about 100 000 c/sec) the prickle and the harmful electrolysis disappear. In this way currents of 1 A and more could be sent through the body. In this case another phenomenon comes to the fore, namely a considerable development of heat in the body.

Heat development of itself has long been an important aid in medical therapy: a local temperature increase is accompanied by a local increase of circulation and a more intense local metabolism, which especially in the case of certain inflammations, rheumatic disorders, etc. may have a favourable effect. In most methods of heat application, however, such as warm baths and packings, massage, irradiation with light or infra-red rays, it is only the outer layer of the body which is heated in the first instance; the subcutaneous layer of fat forms a sort of heat insulator which prevents a direct heat exchange of the more deeply lying parts with their surroundings. This is of course exactly the function of the fat layer in the human body, although it is to prevent the exchange in the opposite direction. The more deeply lying parts of the body can therefore only receive heat by transportation by way of the blood stream. It is clear that no considerable and certainly no well-localized heating of the organs inside the body can be obtained in this way.

Heating by electric current has the important advantage that the heat is developed all along the path of the current, thus also in the internal organs themselves, so that in this case heating effects of quite a different order of magnitude are possible. In the practical application it is then only a question of providing that the electric current follows the desired path through the body so that the desired parts are heated. While it was originally held that for this purpose it was only necessary to apply the electrodes which supply the current to the body at the correct spots, more recent knowledge has shown that this is not sufficient. The part of the body to be treated may in the first instance be regarded as consisting of a very large number of heterogeneous elements connected partly in series and partly in parallel, each of which has its own resistance and its own capacity. When a frequency of for instance 10^6 c/sec (wavelength 300 m) is used, as is the case in diathermy, the reactances of the capacities mentioned are still so small that they exert no influence on the distribution of the current through the parallel elements ; this distribution then depends only upon the resistances, the current chosing the path of least resistance. Now certain components of the body, such as fat, bone, etc., have such a high specific resistance that they practically act as insulators. It is therefore impossible to send a current through parts which are surrounded by fat, although they lie exactly between the applied electrodes: the current does not pass through those parts but around them, *via* muscle tissue and blood stream. It is for this reason that it is impossible to heat the interior of a joint by diathermy.

The situation changes completely, however, when the frequency of the electric current is still further increased to 10^7 c/sec or more (wavelengths of less than 30 m). The reactances of the body elements then become of the same order of magnitude as their resistances and the current distribution is now also influenced to a decided degree by the dielectric constant ε of the various components. Since the value of the dielectric constants of the different components varies less than their conductivity — for most of the body components ε at such high frequencies is of the order of magnitude 80, corresponding to the value of ε for various aqueous solutions — we obtain a much more homogeneous current distribution than at lower frequencies, see *fig. 1*. Thus it is possible actually to send the current through the desired part of the body, and in this way to cause the development of heat at any desired spot.

This is the great significance of the so-called short-wave therapy which has come into practice since 1926. At the same time the very high frequencies used offered another advantage over the lower frequencies used in diathermy. With direct current and with alternating current of not too high frequencies the current must be applied by direct contact on the body, and to prevent a high transition resistance a good contact between the bare electrodes and the skin has to be provided. This makes the treatment difficult and in many cases, such as those of open skin wounds, impossible. At higher frequencies, however, the electrodes can without objection be placed at some distance from the skin. The current is then transmitted *via* the "coupling capacity" between electrode and skin, so that the equivalent circuit shown in *fig. 2* is obtained. Against the advantages of this treatment from a distance there is no other disadvantage than that for the same current through the patient a higher voltage must be applied to the electrodes. The difference is not very important, since when the distance between electrodes and skin is not too great the coupling capacities C_A are of the same order of magnitude as the capacity C formed by the part of the body being treated.

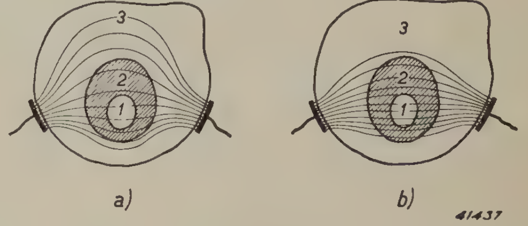

Fig. 1. Current distribution through a part of the body, *a*) at low frequency, *b*) at sufficiently high frequency. *1* is the organ to be heated, *2* an enveloping layer of fat, *3* muscle tissue. At low frequencies the fat has the character of an insulator.

For a correct understanding it is necessary to point out that the historical development of short-wave therapy by no means followed the lines of the discussion given here. It may rather be assumed that the search for effects of the high-frequency alternating current other than those which result from the local development of heat was the main object in view. Thus, for example, it was thought to have been established that after the elapse of some time the operating personnel of powerful radio transmitters suffered from certain bodily phenomena (such as chronic headache) which were ascribed to the effect of the strong electromagnetic alternating fields. Leaving out the question whether or not these suspicions are based upon truth, it must be noted that while the situation of the patient in the "condenser field" in short-wave therapy is quite analogous to that in which one is situated in the vicinity of a transmitting

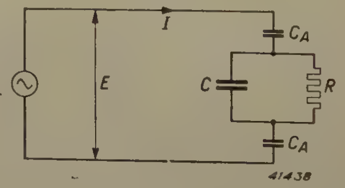

Fig. 2. Equivalent diagram of the circuit which includes the patient in treatment from a distance. The generator furnishes the high-frequency current I which, *via* the coupling capacities C_A existing between the electrodes and the skin, flows through the RC system of the body.

aerial, the specific effects, if they are present, do not play any significant rôle compared with the heat effect. The name "short-wave therapy" is in fact not a very fortunate one, since it emphasizes the wave character, while the fact that waves are also actually emitted by the patient acting as aerial and by the connecting wires is only a subsidiary phenomenon which is undesirable in every respect: it means on the one hand loss of energy and on the other hand it may lead to radio interferences in the neighbourhood. The term "high-frequency diathermy" would therefore be preferable.

The first experiments with short-wave therapy were performed with the help of the oscillations which occur upon the discharge of a condenser battery *via* a spark gap. Such spark generators are still in use at the present time. Short-wave therapy has, however, only been practised on a large scale since it became possible through the development of short-wave transmitting technology, particularly of transmitting valves, to construct entirely noiseless, reliable and easily operated valve generators which deliver a satisfactorily high output. Two such generators which were developed by Philips (types 11 951 and 11 952) will be described in the following.

Choice of the frequency

From the above it follows that the frequency of the generators will be chosen in the region above 10^7 c/sec (wavelength less than 30 m). As to the final choice of the particular frequency, it is more and more customary to work with a wavelength of 6 m. This choice may be explained as follows.

If we consider the diagram of the equivalent circuit in which the patient is included (fig. 2) and calculate the heat Q which is developed per second in the resistance R when an AC voltage of magnitude E and frequency ω is applied to the connections, we find that

$$Q = \frac{E^2}{R\left\{\left(1 + \dfrac{2C}{C_A}\right)^2 + \dfrac{4}{\omega^2 R^2 C_A{}^2}\right\}} \quad . \; . \; . \; . \; . \; (1)$$

It is immediately evident that Q becomes larger, the larger ω is chosen. In order to obtain the greatest possible heat delevopment in the human body, therefore, the frequency must be chosen as high as possible. This conclusion also remains valid when the simple connection in parallel of R and C in fig. 2 is replaced by a complicated circuit of many C's and R's in series and in parallel, which corresponds more to the actual case.

A practical limit is set to the increase of the frequency by two factors. In the first place with all transmitting valves the efficiency decreases when the frequency rises above a certain value. Whereas until a few years ago the limit at which one could work and retain a satisfactory efficiency and output lay at 10 to 30 m, at the present time with newer valves good efficiency can be attained at 6 m and even at 3 m [1]). It is even possible by the employment of magnetron generators [2]) to work with wavelengths of 1 m or less. At such short waves, however, another difficulty occurs, namely the emission of radiation by the connections which lead from the generator to the patient. This radiation can only be restricted by constructing the two electrode cables in the form of a Lecher system, *i.e.* by keeping their distance apart small compared with the wavelength. Considering the very different ways in which the electrodes must be able to be placed, this condition can no longer be satisfied at wavelengths of for instance 3 m. For the same reason, when using wavelengths of less than 1 m, the method has been adopted of radiating the energy on to the patient with parabolic reflectors. Because of these complications the advantage in heat development achieved by the increase in the frequency is again endangered, so that for the present it is best not to choose a frequency higher than 5 × 10^7 c/sec (wavelength 6 m). This wave length is the one most commonly used for short-wave therapy at the present time, and it has been taken as a basis for the Philips generators.

Here again it must be noted that the historical development has not followed this argument as a whole. A group of investigators held the opinion that for heating different parts of the body different frequencies should be chosen, in order to obtain in each case the largest possible heat development[3]). This conception originated from the fact that when equation (1) or one similar to it set up for constant current is partially differentiated with respect to R and $\mathrm{d}Q/\mathrm{d}R$ is set equal to zero, a relation between R, C, C_A and ω is found for which Q has a maximum. This means, however, — as Daan [4]) pointed out — nothing more than that at a given frequency a certain relation between R and C is the most favourable. If on the other hand R and C must be considered as given, as is always the case in practice, the best choice of ω is independent of R and C, as explained above.

The high-frequency output required

The power which must be supplied to the patient depends upon the size of the part of the body to be heated. It must be kept in mind that the energy must be supplied at a certain rate if any appreciable increase in temperature is to take place. The circulatory system inside the body and the perspiration of the skin have a very strong heat-regulating action, *i.e.* they combat local changes in temperature. The heating of a part of the body must overcome these equalizing influences. On the other hand the heating may not take place too quickly, since the body reacts against too rapid disturbances of the heat equilibrium by severe sensations of pain. If for example it is desired to give a temperature increase of 1 to 1.5°C to the chest or pelvis of a patient, a heating time of about 5 min. is found to be necessary, while a power of about 300 W must be supplied to the patient. This is about the highest power which will be needed for the treatment of a single object. Only in the case of the excitation of artificial fever, which was formerly carried out by expressly infecting with malaria, and where the purpose is to heat the whole body several degrees centigrade, are higher powers required, namely 700 to 800 W. Since, however, this treatment is quite rare, it is better to use two separate 400 W apparatus than a heavy apparatus constructed especially for that purpose, also because a 400 W apparatus can more easily be built to make it transportable — a requirement which is always made of apparatus to be used in a hospital. For these reasons the Philips apparatus 11 952 is built for an output of 400 W. This apparatus has the form of a desk on wheels, see *fig. 3*. In order also to be able to treat bedridden patients at home a second type (11 951) has also been developed which with an effective output of 150 W weighs only 28 kg. In this case therefore some of the heaviest treatments are impossible, but the apparatus can easily be transported in a motor car (*fig. 4*).

---

[1]) See Philips Techn. Rev. **6**, 253, 1941.

[2]) See for example Philips Techn. Rev. **4**, 189, 1939.

[3]) J. Pätzold, Z. Hochfrequenztechnik **36**, 85, 1930.

[4]) A. Daan, dissertation, Amsterdam 1938.

The above refers only to the maximum output for which the generator must be designed. On the subject of dosage, *i.e.* the regulation and control of the power applied to the patient, some remarks will be made in the following.

Connections of the high-frequency generator

Diagram showing the principle

In principle a generator for short-wave therapy is nothing else but a short-wave radio transmitter. It can, however, be made considerably simpler than a radio transmitter, due to the fact that for therapy it is unnecessary to make strict requirements as to the constancy of the frequency and the amplitude of the oscillation generated. Since an especially constant frequency is not required, driving stages with crystals or the like can be omitted and, even for relatively high outputs, a self-oscillating connection with only one stage can be used. Owing to the fact that a stable amplitude is not required (absence of modulation), the generator can be fed with AC voltage, so that no rectifiers are needed for the supply. It is in fact only these simplifications which make it possible to design a unit which is acceptable economically to the existing circle of users.

The diagram of the fundamental connections of the generators 11 951 and 11 952 is reproduced in *fig. 5a*. The anode circuit, which is tuned to a frequency of 5×10^7 c/sec, is formed by the anode-cathode capacities of the two transmitting valves and the two coils L_a', L_a''. The connections show some similarity to the push-pull connections often used in radio technology (fig. 5*b*). They differ from the latter, however, in that the two valves oscillate in turn, each valve oscillating only in the half period of the low-frequency supply voltage in which its anode is positive, while in the case of true

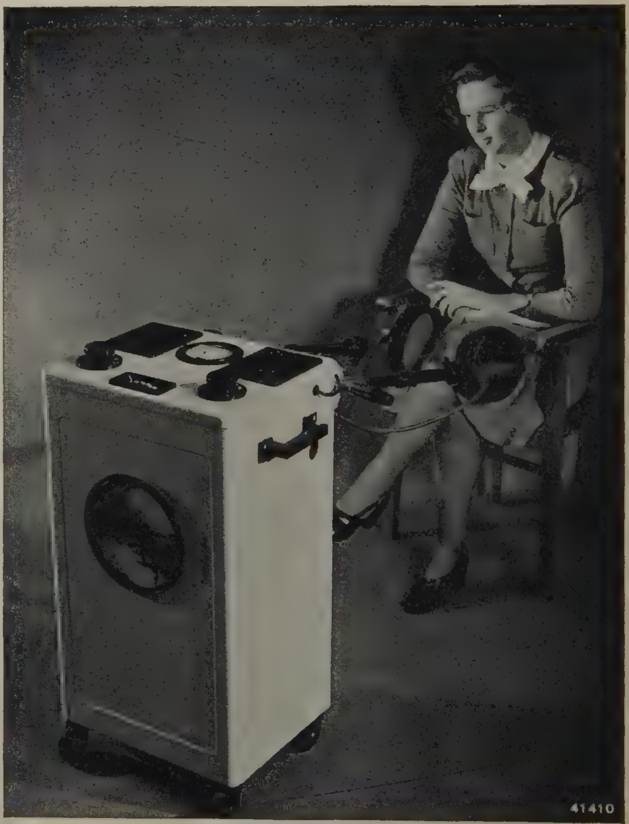

Fig. 3. Generator for short-wave therapy, type 11 952, in use (treatment of knee joint). This apparatus in the form of a desk on wheels can deliver a maximum high-frequency output (wavelength 5 m) to the patient. Besides the knob for switching on (right) there is no other part to be operated by the medical personnel but the knob for adjusting the desired output (left), which is read off on the meter, and the disc for tuning the circuit with the patient (centre), which is checked on the same meter. Dimensions 46 × 37 × 87 cm, weight 62.5 kg. Two electrodes in the form of plates are fastened to the apparatus by means of two arms which can be turned in all directions.

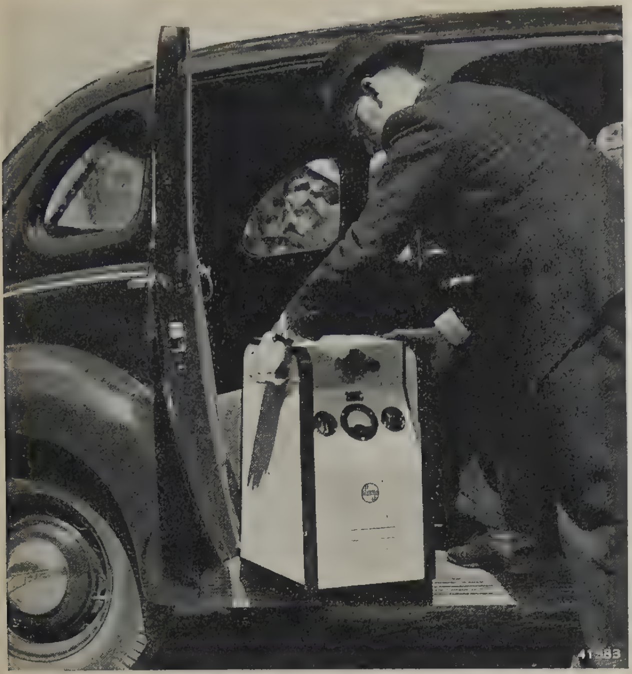

Fig. 4. Generator for short-wave therapy, type 11 951. The apparatus weighs only 28 kg and can easily be transported in a motor-car. The maximum effective high-frequency output amounts to 150 W.

push-pull connections both valves oscillate continuously, see *fig. 6a*. Nevertheless, even in these halves of the period during which the "other" valve is generating, the full high-frequency voltage acts on both of the anode coils, due to their mutual induction. The high-frequency voltage on the complete circuit has, therefore, at every moment an amplitude which in the most favourable case (full modulation of the voltage) is four times as large as the momentary value of the low-frequency supply voltage, see fig. 6*d*.

The desired power can of itself also be excited without difficulty with a single valve. However, by using two valves each of lower output and connecting them as in fig. 5*a*, several important advantages are obtained. In the first place the supply transformer is symmetrically loaded, so that the primary current is about twice as small (cos $\varphi \approx 1$, while with unsymmetrical loading cos $\varphi \approx 0.5$), and thus the transformer can be made smaller. Furthermore, the symmetry of the high-frequency oscillation circuit facilitates the oscillation and diminishes the reaction of the circuit in which the patient is included on the generator. Thanks to the symmetrical connections there is also less chance of high-frequency voltages being transmitted to the supply network. Finally it must be noted that when using a single valve intervals occur during which there is no high-frequency voltage present, as may be seen in fig. 6*b* or *c*. On the other hand when using two valves the high-frequency voltage is more nearly continuously present. Some patients seem to find the latter situation more pleasant.

The circuit L_pC_p in which the patient is included is coupled inductively with L_a'—L_a'', so that the low-frequency anode supply voltage, which in the case of valves for several hundred watts must normally amount to several thousand volts, cannot act upon the patient. For absolute safety the point of symmetry of L_pC_p is furthermore connected with earth *via* a high-frequency choking coil (not shown in fig. 5*a*). The circuit L_pC_p is tuned to the generator frequency for every case by means of a rotating condenser.

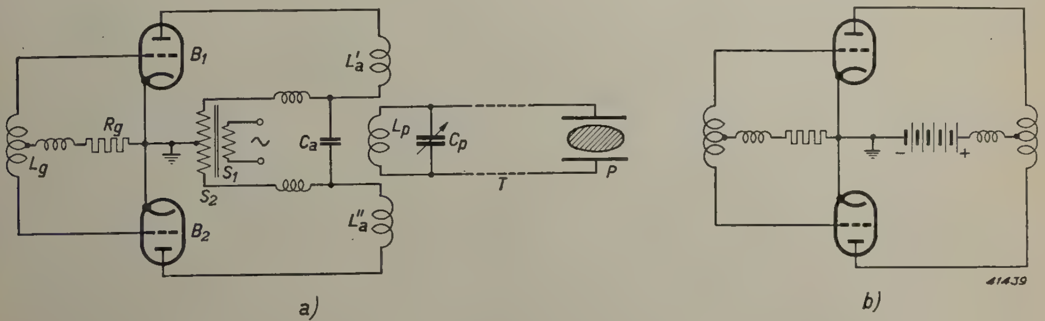

Fig. 5. *a*) Diagram showing the principle of the Philips generators for short-wave therapy. Since the requirements as to constancy of the frequency and the amplitude of the oscillation obtained are not severe, the generator is built up of a single self-generating stage, and as anode voltage simply a 50 c/sec AC voltage is taken which is delivered by the supply transformer S_1, S_2. The symmetrical connection of two valves (B_1, B_2) here employed has various advantages over connections with only one valve. The circuit in which the patient P is included *via* the supply lines T and which is tuned to the wavelength 6 m with the rotating condenser C_p for each case treated, is coupled with the divided anode coil L'_a—L''_a by the coil L_p. C_a is a coupling condenser, L_g grid coil, R_g grid resistance.
b) Normal push-pull connection of two valves.

While the construction of a medical shortwave apparatus is on the one hand simpler than that of a radio transmitter, thanks to the lower requirements of constancy of frequency and amplitude of the oscillations obtained, there are on the other hand complications which are not encountered in a radio transmitter: 1) the loading impedance is different for every case treated, 2) it must be possible to vary the output within wide limits (practically by a factor 10).

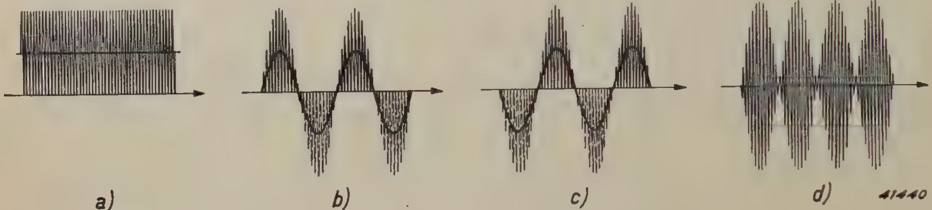

Fig. 6. Form of the resultant anode voltage
a) of each of the two valves in fig. 5*b*,
b) of valve B_1 in fig. 5*a*,
c) of valve B_2 in fig. 5*a*,
d) of the two valves in fig. 5*a* together.
The anode DC voltage and low-frequency anode supply voltage, are indicated by heavy lines. The oscillations indicated by thin broken lines in *b* and *c* are due to induction in the corresponding half of the anode coil during the "wrong" period of the supply voltage.

Matching of the circuit in which the patient is included

As to the loading impedance, the loading of the anode circuit should be matched for each patient by making the coupling between L_a and L_p more or less tight. With the technically unskilled operators of the medical apparatus, however, it is not easily possible to have them make this adjustment, not so much because of the fact that the coupling requires too great precision as far as the magnitude of the output is concerned (with self-generating connections the opposite is true), but because with too tight a coupling an undesired reaction of the circuit with the patient on the generator may occur. Therefore in the apparatus the coupling between L_a and L_p is adjusted permanently in such a way that for the heaviest cases of loading occurring in practice (smallest impedance in parallel with L_pC_p) the correct matching is obtained, while with larger parallel impedance the coupling is actually too loose to make the generator deliver the maximum output. This is no objection since with larger parallel impedances, *i.e.* smaller parts of the body, the maximum output is not applied, in connection with the above-mentioned requirement that the heating up may not take place too quickly.

This brings us to the second complication, the regulation of the power supplied to the patient and the problem of measuring this power.

Regulation of the output

If one begins with the maximum output, the output can be reduced either by decreasing the heating current, increasing the grid resistance or decreasing the anode voltage. The last method is the most economical one, since in that case at lower outputs no unnecessary losses occur in the apparatus. This method has been applied in the larger of the two generators here described, and the regulation is made by varying the primary voltage of the supply transformer. In this way it is of course impossible to take the heating current of the valves simply from an extra secondary winding of the supply transformer, and a separate heating current transformer must be used. Against this disadvantage, however, is the fact that the regulation at the primary side makes possible a simple measurement of the effective output. We shall explain this briefly.

For the correct adjustment of the output sometimes the only criterion which is used is that the patient feels a "pleasant warmth". It is obvious that this is not a very reliable method, if only because in many cases (for instance frost bite and various neurological disorders) the sensitivity to heat is disturbed at the spot to be treated. It is impossible to include a direct-reading instrument in the circuit with the patient, since no wattmeter suitable for these high frequencies exists. A useful method of determining the high-frequency output consists in substituting for the patient an electric lamp of about the same resistance, allowing the light flux obtained to fall upon a luxmeter, and afterwards reproducing the indication of the luxmeter by supplying the lamp with direct current or low-frequency alternating current with wattmeter connected in series 5)). This method is indeed applied for control in the factory, but it is, too elaborate for use by the doctor.

In this method of regulating the output variation of the primary supply voltage with unvarying grid resistance and heating current, the ratio between the "DC voltage" (actually: low-frequency DC voltage) and the high-frequency AC voltage remains unaltered, *i.e.* the modulation of the generator valve always remains the same. By this means and by means of measures for the limitation of losses, which will be briefly discussed below, it is possible to obtain a practically constant efficiency of the generator for all values of the output chosen. This means that the ratio between the power supplied to the patient and the power taken up from the mains is constant, and we may therefore determine the former simply by measuring the latter, which may be done with a normal wattmeter preceding the supply transformer.

Several particulars of the construction

The apparatus 11 952 contains two triodes type TB 2/200, which type has a permissible dissipation of 130 W, and at 6 m wavelength in the connections described it can develop a high-frequency output of 200 W6)). In the smaller apparatus 11 951 four small triodes (type no. 29 933) are used and connected two by two in parallel.

With full load the valves of the apparatus 11 952 operate with an efficiency of $200/(200+130) \approx 60$ percent. The total efficiency of the apparatus is of course lower, due to all kinds of losses. These losses are partially proportional to the power supplied to the patient and partially depend upon still other factors. The latter, among which are losses by radiation, were limited as far as possible in order to ensure the proportionality required for the measurement of the output described between the power supplied to the patient and the power taken from the mains. The remaining deviations from this proportionality cause in the most unfavourable case a measuring error of ± 15 percent, which for medical application already constitutes a very satisfactory accuracy compared with the inaccuracy which had hitherto prevailed.

Moreover, those losses which do not affect these measurements, since they are proportional to the effective output, should also be restricted for the sake of economical operation. Such losses occur for example at the grids in the circuit represented in *fig. 5a*. When the anode of one valve is positive the oscillations excited by that valve act also on the grid of the other valve, due to back-coupling and induction in the grid coil. While of course the anode of the latter valve is negative at that moment, grid current can, nevertheless, flow in the positive peaks of the grid voltage, and this grid current results in an increased grid dissipation. It has been found possible to decrease this effect considerably by modifying the connections slightly, see *fig. 7*. Each of the two grids is connected with a tap on its "own" half of the supply winding S_2, which tap is so chosen that in the period of negative anode voltage the grid receives a sufficiently large negative bias to prevent grid currents due to the induced high-frequency voltage. Since the grids of the two valves must be at different low-frequency voltages, the grid coil L_g is now interrupted by a condenser, just as was previously done with the anode coil L_a. In the period of positive anode voltage the extra grid bias now becomes positive. In order to prevent the flow of very large grid currents due to this cause, which would again neutralize the desired effect, the grid resistances R'_g, R''_g are increased so much that the grid current takes on a normal relation to the anode current. In this way it has been found possible to obtain an average total efficiency of the apparatus of 45 percent.

The various details mentioned here can be recognized in the photograph of *fig. 8*. The construction of the coupling condenser between the two halves of the anode coil is striking. This condenser is at the full voltage of the supply winding (several thousand volts) and carries the high-frequency current of the anode circuit (order of magnitude 10 A). It is therefore made of a special ceramic material which combines low high-frequency losses with high voltage loadability.

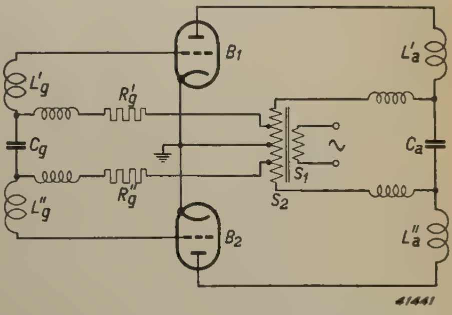

Fig. 7. In order to limit the grid dissipation, the grid resistances R'_g and R''_g, for each of the valves in fig. 5a are connected with a certain tap on the corresponding half of the winding of the supply transformer.

---

5) *Cf.* also Philips Techn. Rev. 6, 215, 1941.
6) See the article referred to in footnote 1).

Practical application of the apparatus

Since the apparatus consume only relatively little current — the larger apparatus a maximum of 4.5 A at 220 V —, they may be connected to any ordinary wall contact which is protected by 6 A fuses. The necessary manipulations are limited to the application of the electrodes to the patient, the tuning of the circuit with the patient on the wavelength of the generator and the adjustment of the desired output. The tuning is easily performed by means of a rotating condenser with fine regulation (in fig. 3 the tooth-edged disc at the front of the apparatus) and in the case of the apparatus 11 952 it is controlled in a simple way due to the fact that the wattmeter for the measurement of the power delivered gives a maximum deflection at the correct tuning. For control of the tuning a gas-discharge tube can also be used. The tube is placed in the vicinity of the patient or the electrodes and gives the most intense illumination at the correct tuning.

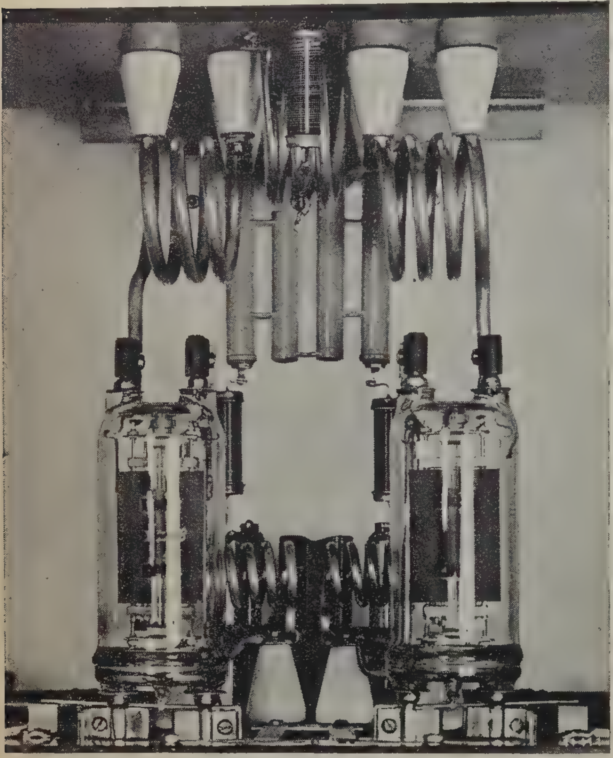

Fig. 8. Part of the inside of the generator 11 952. Above the two valves are placed the two halves of the anode coil; partly between and partly above these coils is the coil of the circuit with the patient. It may be seen that the coupling is only very loose. Behind the coils hangs the coupling condenser of the anode coil, constructed of special ceramic material. Between the valves may be seen the grid coil, which is also divided into two halves, and behind it the corresponding coupling condenser.

As to the application of the electrodes, it need only be noted that the doctor in charge must have some insight into the expected course of the lines of force between the electrodes. By a correct choice of shape, size, position and distance from the skin he can then cause the heat to be developed as nearly as possible in the desired spots. The diagrams in *fig. 9 a-c* and *fig. 10* give a few examples, while the photograph of fig. 3 shows how the electrodes are fastened to the generator and how in a given case the arrangements can be made. A more detailed discussion of the various forms of electrodes in use and of the methods of treatment, however, falls outside the scope of this article.

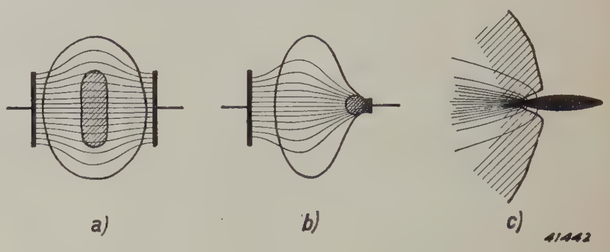

Fig. 9. By giving the electrodes a suitable form and placing them in a suitable position the desired localization of the heat development in the body can be realized. *a*) Uniform heating of an extensive region with two plate electrodes. *b*) Concentration of the heat development in a small region (furuncle) with a small plate electrode. *c*) So-called electro-surgery: by connecting the knife which makes the incision in an operation to one pole of the short-wave apparatus, while the other electrode is placed arbitrarily, such a strong concentration of lines of force occurs at the point where the incision is made that the blood at that point immediately coagulates from the heat (bloodless operation). The cutting itself, *i.e.* the severing of the tissue, also takes place much more easily.

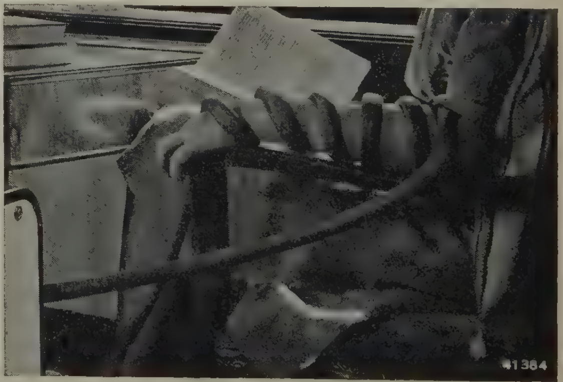

Fig. 10. In order to heat an arm or a leg uniformly, the member in question is not placed between two electrodes, but a coil connected with the two poles is wound around it (method of Kowarschik).

THE VIBRATION OF CONTACT SPRINGS

by J. A. HARINGX.

621.3.066.6

For the satisfactory operation of relay contacts it is found that it is important that the closing of the relay should take place very gradually, because otherwise the contact shows a tendency to rebound repeatedly before it remains definitely closed. The occurrence or absence of an impulse is found to depend upon the resonance frequency of the contact spring. In this article the resonance frequencies are calculated which must be chosen in order to obtain a velocity of closing equal to zero. It is furthermore shown that with only a slight deviation from these resonance frequencies the velocity of closing will increase appreciably.

In different kinds of electrical apparatus, such as for example the electric bell, relay contacts are used which are opened and closed periodically at a fairly high frequency. For the sarisfactory functioning of the mechanism it is very important that the opening and closing should occur exactly at the moments fixed for the structure in question, and, moreover, not only when the apparatus is new, but also after long use.

One of the phenomena which must be considered is the elastic vibration of the contact springs. In particular, it may sometimes be observed that after each mutual contact between the two parts of the contact, the spring continues to vibrate so that the electrical connection is broken several times before it is finally made. Since a spark occurs upon each interruption of the current circuit, it is obvious that such a vibrating contact burns through much more rapidly than a contact which is only broken once per period. Moreover, disturbing high-frequency phenomena occur as a result of these sparks.

The problem of the vibration of contact springs has already been discussed in this periodical in connection with a vibrator for the conversion of direct current into alternating current [1]). In that case the striking together of the two parts of the contact was compared with the collision between two billiard balls, in which case it may also be observed that the contact which occurs is immediately broken again under the influence of the elastic forces. Under special circumstances, however, the contact between the moving ball and the stationary ball is maintained. This case may occur when a third ball is present touching the stationary ball. The impulse of the moving ball can then be transmitted through the stationary ball to the third ball, so that the latter rolls away while the first two balls remain in contact.

In the article referred to above a description is given of the method by which this principle of the third billiard ball can be usefully employed in relay contacts. In this way the two colliding metal parts of the contact are made to remain in contact without vibration, while the impulse is transmitted to the third mass, which is elastically attached. This latter mass can be knocked away without, however, any harmful results, since the broken contact between the second and third parts of the contact does not constitute part of the circuit.

Since this method of combating vibration in contacts is quite complicated, it seemed to us important to find out whether it was not possible to avoid the necessity of such measures. By a suitable choice of the characteristic frequency of the vibrating contact spring it has been found possible to cause the two parts of the contact to come together very gradually, so that no shock at all occurs. Special measures of dissipating the impulse of the collision thus become superfluous.

Mathematical formulation of the problem

The relay contact can be represented by a point of mass *1* at the end of a spring (see *fig. 1*).

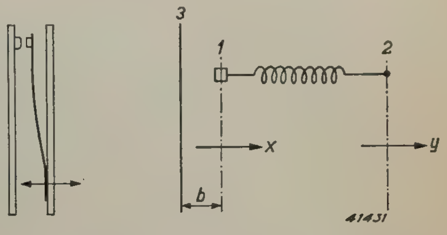

Fig. 1. Principle and mechanical model of a spring contact.

The other end (*2*) of this spring makes a sinusoidal motion according to the formula

$$y = a \cos \omega t \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The mass point *1* will then also execute a sinusoidal motion, provided the amplitude A of this motion remains smaller than the distance b between the stop *3* and the position of rest. If m is the mass of the vibrating part of the contact and c the stiffness of the spring, *i.e.* the force per unit of elongation, the following is the differential equation of this motion:

---

[1]) Philips techn. Rev. 6, 342, 1941.

$$m\frac{d^2x}{dt^2} = c\,(y-x),$$

or if we substitute for y the value according to (1)

$$\frac{d^2x}{dt^2} + \frac{c}{m}\,x = \frac{c}{m}\,a\cos\omega t.$$

The general solution of this is

$$x = A\cos\omega t + B\cos\omega t_0 + C\sin\omega_0 t, \quad . \; . \; . \quad (2)$$

where the characteristic frequency $\omega_0 = \sqrt{c/m}$

and $$A = \frac{a}{1-\omega^2/\omega_0{}^2} \quad . \; . \; . \; . \; . \; . \; . \; . \; . \; . \; . \; . \quad (3)$$

The term $A\cos\omega t$ represents the forced vibration of the mass point, while the other two terms of equation (2) describe the free vibration of the mass point. Let us now examine what happens when the deflection of the mass point is limited by the stop *3*. In a certain part of the period the mass point will then have the deflection $-b$ and will lie against the stop with a certain pressure. The motion of the mass point will only begin again when the pressure ceases to exist, in other words when the compressed spring has just recovered its original length. Then evidently $y = -b$, so that for the moment t_1 of the beginning of the motion we may write

$$a\cos\omega t_1 = -b \; . \; . \; . \; . \; . \; . \; . \; . \; . \quad (4)$$

The motion of the mass point must therefore be such that at the moment $t = t_1$, there is a deflection $-b$ and the velocity is zero. These conditions are sufficient to calculate B and C, so that the mechanical problem is hereby solved in principle.

A possible solution is reproduced in *fig. 2*. The motion of point *2* may be seen and the forced vibration of point *1*, as well as the free vibration which must be added to fulfil the initial conditions. At the same time the resultant motion of the mass point is reproduced. t_1 is the moment at which the movement begins; t_2 the moment at which the contact is again closed. It may be seen that the velocity of the motion (slope of the curve) at the moment of closing has a very discontinuous character, which means that a strong impulse will occur upon closing. Our object of procuring a gradual coming together of the parts of the contact is thus not realized in this case.

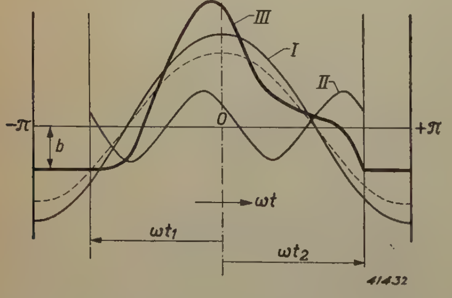

Fig. 2. Possible motion of the end of a contact spring whose other end moves sinusoidally according to the broken-line curve. The curves *I*, *II*, *III* indicate respectively the forced vibration, the free vibration and the resultant motion.

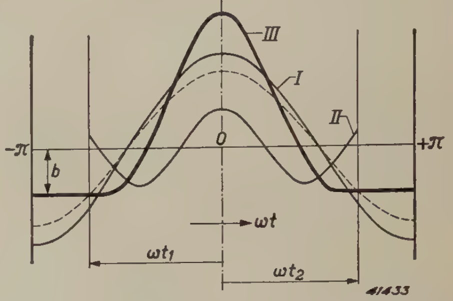

Fig. 3. Motion of the end of a contact spring whose resonance frequency is so chosen that the closing velocity is zero.

In order to obtain improvement it is now sufficient merely to choose the characteristic frequency of the contact spring somewhat lower, so that the diagram of *fig. 3* is obtained. The forced vibration and the free vibration of the mass point are now both symmetrical with respect to the dot-dash line, which represents the moment $t = 0$. The resultant motion of the mass point is thus also symmetrical with respect to this line, and since the motion begins with a velocity zero, it will also end with a velocity zero, so that the desired impulse-free contact is obtained.

The symmetry condition can be formulated mathematically by saying that the free vibration is a pure cosine function; the term $C\sin\omega_0 t$ in equation (2) must therefore dissappear [2]). Since C depends also upon ω_0, by a suitable choice of the resonance frequency of the spring provision can be made to satisfy this condition.

Calculation of the desired resonance frequency

As we have seen, the constants B and C of the motion of the contact piece given in equation (2) are determined by the conditions that for $t = t_1$ there should be a deflection $x = -b$ and a velocity $dx/dt = 0$. Thus in formulae:

$$A\cos\omega t_1 + B\cos\omega_0 t_1 + C\sin\omega_0 t_1 = -b,$$
$$-A\omega\sin\omega t_1 - B\omega_0\sin\omega_0 t_1 + C\omega_0\cos\omega_0 t_1 = 0.$$

---

[2]) Actually it has only been proved that the disappearance of the term $C\sin\omega_0 t$ is sufficient for securing a closing velocity of zero. To complete the mathematical reasoning it should therefore also be proved that impulse-free closing cannot be secured in any other way, so that the condition $C = 0$ is actually necessary. We shall not, however, go into that here.

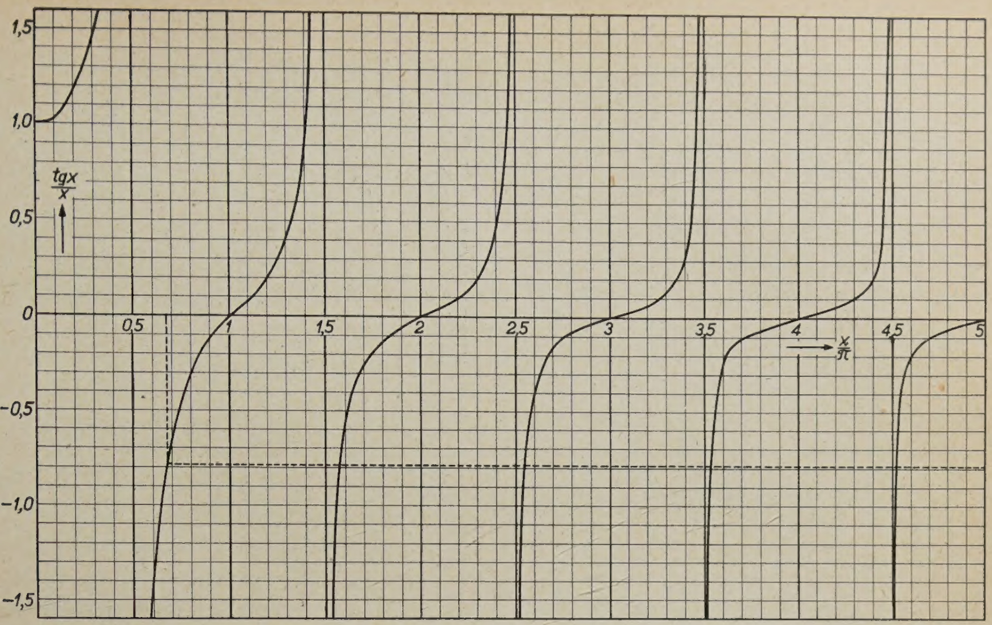

Fig. 4. The function tan x/x which is used to solve the transcendental equation

$$\frac{\tan kx}{kx} = \frac{\tan x}{x}$$

for k. From the figure it may, for example, be deduced that $x/\pi = {}^2/_3$ corresponds to $kx/\pi = 1.57$; 2.59; 3.54; 4.52; etc., and therefore to $k = 2.35$; 3.88; 5.32; 6.78 etc.

By substituting in these formulae the following values from equations (4) and (3):

$$b = -A\ (1-\omega^2/\omega_0{}^2)\cos\omega t_1,$$

one finds

$$\left.\begin{aligned} B\cos\omega_0 t_1 + C\sin\omega_0 t_1 &= -\frac{\omega^2}{\omega^2_0}A\cos\omega t_1,\\ -B\sin\omega_0 t_1 + C\cos\omega_0 t_1 &= \frac{\omega}{\omega_0}A\sin\omega t_1. \end{aligned}\right\} \quad (5)$$

If now, in agreement with the condition for impulse-free closing, we make the condition that $C = 0$, we may divide the two equations (5) by each other and in that way find the condition that

$$\tan\omega_0 t_1 = \frac{\omega_0}{\omega}\tan\omega t_1 \quad . \ . \ . \ . \ . \ . \quad (6)$$

In order to draw practical conclusions from this equation we shall pass over to new variables which are adapted to the form of the technical problem. The function of the mechanism usually prescribes that the contact shall be open during a certain part of the period lying between 50 and 100 percent. If we let this fraction equal α, then according to fig. 3

$$\omega t_1 = -\pi\alpha.$$

If, moreover, we set ω_0/ω equal to k, (6) may be written in the form

$$\frac{\tan k\pi\alpha}{k\pi\alpha} = \frac{\tan\pi\alpha}{\pi\alpha}$$

and we obtain in this way a relation with whose help k can easily be determined for every prescribed value of α. To do this it is only necessary to draw the function tan x/x and to find the values of x for which the ordinate is equal to tan $\pi\alpha/\pi\alpha$ (*see fig. 4*).

If α is chosen equal to 0.5, which means that the relay is open for one half of the period, tan $\pi\alpha/\pi\alpha$ is infinitely large, and the same must be true of tan $k\pi\alpha/k\pi\alpha$. From this for k all the odd numbers follow:

$$k = \frac{\omega_0}{\omega} = 1,\ 3,\ 5,\ 7,\ 9,\ \ldots$$

In the limiting case where $\alpha = 1$, thus for a relay which is open during the whole period, tan $\pi\alpha/\pi\alpha = 0$. The tangent of $k\pi\alpha$ (with $\alpha = 1$) must therefore also be zero, and from this it follows that

$$k = 1,\ 2,\ 3,\ \ldots$$

Thus if α is allowed to increase continuously from 0.5 to 1, k must pass from the series of odd numbers to the series of all whole numbers. The transition conditions are easily constructed with the help of fig. 4, as may be seen from the example given. Finally in *fig. 5* the result of this graphic calculation is given, namely the ratio $\frac{\omega_0}{\omega}$ as a function of α for the case where the contact is closed with no impulse.

Tolerances for the resonance frequency

For the practical application of the method discussed for obtaining a closing velocity of the contact equal to zero it is not only important to know the suitable resonance frequency of the contact spring, but at the same time one must know how accurately this frequency must be adjusted in order to obtain the desired effect.

If we consider as example the case where $a = 0.5$, we see that the forced vibration $A \cos \omega t$ has the maximum velocity just at the moments t_1 and t_2 at which the contact opens or closes. For the free vibration $B \cos \omega_0 t$, in the case of a closing velocity of zero, the same is true, since ω_0 is an odd multiple of ω. The coefficient B has a value such that the velocities of the forced vibration and of the free vibration compensate each other exactly for t_1 and t_2.

If ω_0 is now allowed to vary, the two velocities will continue to compensate each other for the initial moment t_1 of the motion, but for the moment t_2 this is in general no longer the case.

If for example ω_0 is an even multiple of ω, the two velocities will reinforce each other as much as possible, so that the resultant velocity is twice as great as that of the forced vibration alone. If the resulting closing velocity v is drawn as a function of ω_0/ω, a wavelike curve is therefore obtained which has maxima for all the odd whole numbers. The minima are zero, while the maximum velocity is twice as high as that of the forced vibration alone. The curve derived qualitatively in this way is reproduced in *fig. 6.*

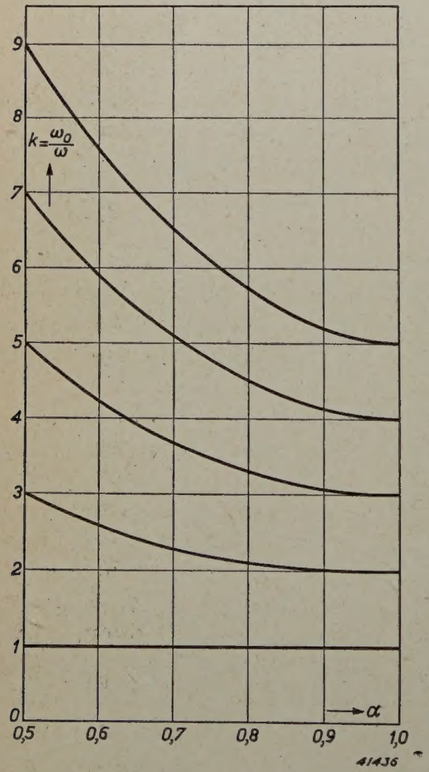

Fig. 5. The behaviour of $\omega_0/\omega = k$ as a function of a for the case of impulse-free closing.

The horizontal line at the height 1 indicates the amplitude of the velocity of the fixed end of the spring. Thus there can only be a question of a decrease in the closing velocity compared with the velocity of the driving motion at those values of k for which the wave curve lies below this horizontal line. As may be seen, this is the case only in sufficiently close proximity to the odd whole numbers; the maximum difference which may be permitted amounts to about $\Delta k = \pm 0.5$.

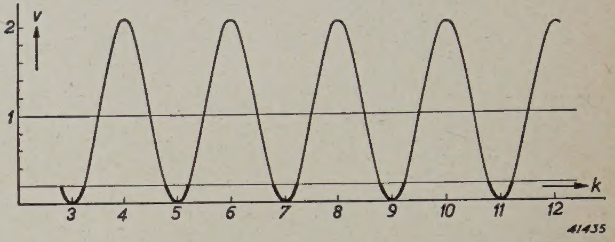

Fig. 6. Closing velocity v as a function of $\frac{\omega_0}{\omega}$ for $a = 1/2$. The ratio of the closing velocity to the velocity which would hold for a rigid connection instead of a spring is plotted as ordinate.

If the condition is made that the decrease in the closing velocity should amount at least to a factor 5, the permissible deviation Δk is reduced to 0.2. This means that the adjustment to the correct resonance frequency is quite critical, since with the relay constructions in practical use values of k lying in the neighbourhood of 7 or 9 are quickly reached. The permissible deviation would then amount to only about 3 percent.

In the foregoing we have been assuming that the fixed end of the spring executes a fixed motion. Usually, however, this motion itself is again a free vibration of a mass M attached to some kind of elastic system. A closer consideration of the more complicated case of a coupled vibrating system shows that in general ω must be set equal to the characteristic vibration of the main mass M, while ω_0 can be calculated with the help of the stiffness of the whole elastic system, considering the mass M as stationary.

When the desired resonance frequency of the relay spring has been found there remains the practical problem of the method of adjusting the resonance frequency accurately. We shall not go into that at this point, since in every case it depends upon the special construction. It may only be stated that in the construction of the vibrator for the conversion of direct current into alternating current mentioned at the beginning of this article, the correct setting of the contact spring was actually found to produce the desired result.

SUMMARY OF RECENT SCIENTIFIC PUBLICATIONS ISSUED BY N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these publications will be gladly supplied free of charge upon request to the Administration of the Natuurkundig Laboratorium, Kastanjelaan, Eindhoven. Reprints of any publications marked with an asterik, however, are not available in sufficient number for distribution.

1570: E. J. W. Verwey: The electrical double layer of oxidic substances especially in non-aqueous media (Rec. trav. chim. Pays Bas **60**, 625-633, July-August 1941).

The electrical double layer on the surface of particles of oxides is studied : 1) by the examination of suspensions of these substances in water, alcohols, acetone or solutions of electrolytes in these liquids and 2) by electro-endosmosis through membranes in these liquids. The double layer is caused by ions originating from the materials examined. The ions of hydrogen and of the hydroxyl group determine the potential. The zero point of the charge is very closely related to the size of the particles. Just as in the case of silver iodide, large particles have a stronger negative charge than small particles.

1571: J. A. M. van Liempt: Eine einfache Methode zur Bestimmung der Diffusionskonstante von Metallen. (A simple method of determining the diffusion constants of metals) (Rec. trav. chim. Pays Bas **60**, 634—639, July-August 1941).

A simple method is developed for determining the diffusion constant in metals. A thin wire or strip of one metal is embedded in a piece of a second less precious metal by metal-ceramic means. The duration and temperature of the heating are determined at which the diffusion has progressed so far that the limit of resistance is just reached over the whole section of the wire or strip.

1572: J. L. Snoek: Concerning the internal damping of ferromagnetic substances. (Ned. T. Natuurk. **8**, 177-189. April 1941).

Internal damping phenomena in iron and other ferro-magnetic materials are discussed. The damping of torsional vibrations in iron and nickel wires is almost exclusively due to ferromagnetic hysteresis; external changes in the magnetisation, however, are not observed. The hysteresis damping appears to be proportional to the amplitude of the vibration. In longitudinally vibrating ferro-magnetic rods considerable changes in the induction can be observed when the magnetostriction is great, the material is properly annealed and an external magnetic field of a suitable strength is applied. When, moreover, the frequency is so chosen that the eddy current losses in the rod are the maximum, exceptionally large values are obtained for the logarithmic decrement, as observed experimentally. In iron and iron alloys containing small quantities of carbon or nitrogen damping phenomena of a certain kind are observed. This is purely an after-effect and shows no hysteresis. The great temperature-dependency observed leads one to suppose that diffusion of carbon or nitrogen through the iron lattice plays a part. In cold-treated samples after-effect phenomena are found in another temperature range. This is complementary to what is observed with well annealed specimens.

1573: Balth. van der Pol: Two little known experiments. (Ned. T. Natuurk. **8**, 390-393, September 1941).

At a meeting of the Netherlands Physical Society at Eindhoven some striking, subjective colour phenomena were demonstrated with a Benham optical disc, such phenomena occurring in spite of the fact that the disc consists only of black and white parts. These colour phenomena also occur when monochromatic yellow sodium light is used. Further, a demonstration was given with two constructions of a hot air engine in its most primitive form (Griffiths' experiment).

1574: M. J. Druyvesteyn and J. L. Meyering: An approximate calculation of the thermal expansion of solids I (Physica **8**, 851-861, September 1941).

The expansion coefficient for a solid body is practically equal to the third differential quotient of the lattice energy in length. For the alkali metal an expansion coefficient is calculated which agrees very well with the experimental values.